may 1961
the
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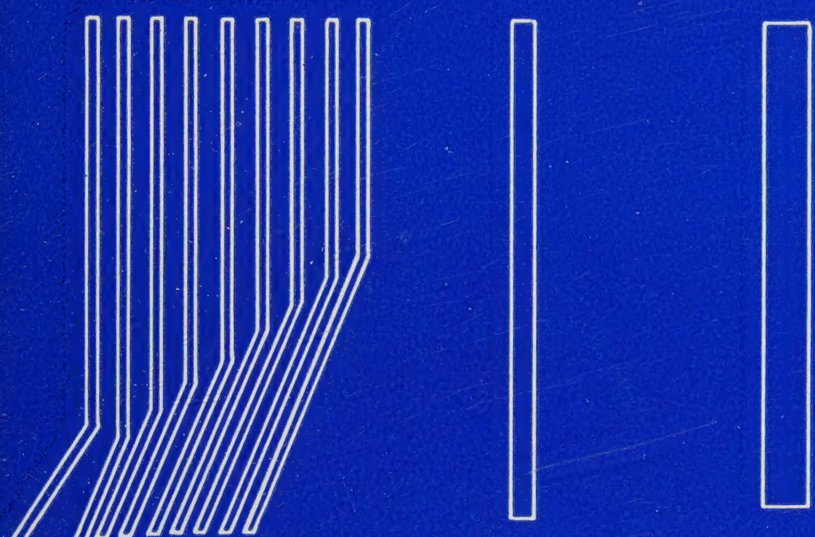
engineers

Proceedings of the IRE

in this issue

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VOLUME 49, NUMBER 5, PART I

May, 1961

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Proceedings of the IRE®

contents

COVER

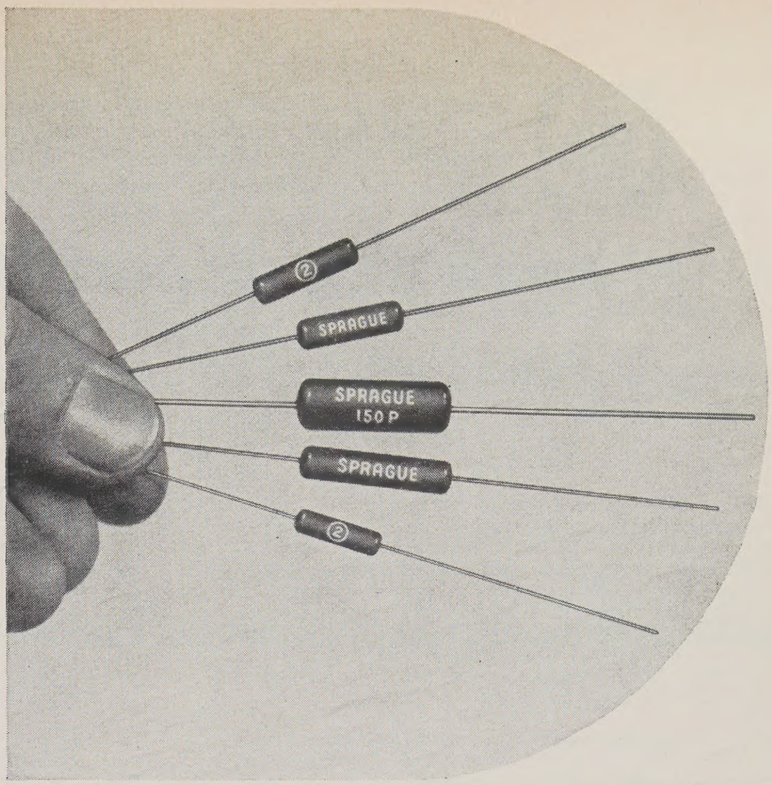
By stacking the anodes of a traveling-wave tube in the unusual array shown on the cover, it is possible to reduce the beam noise to a level formerly thought to be impossible. This is but one of several important techniques for obtaining very-low-noise amplification which are reviewed on page 880.

Proceedings of the IRE®

continued

PROCEEDINGS OF THE IRE, published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N. Y. Manuscripts should be submitted in triplicate to the Editorial Department. Correspondence column items should not exceed four double-spaced pages (illustrations count as one-half page each). Responsibility for contents of papers published rests upon the authors, and not the IRE or its members. All republication rights, including translations, are reserved by the IRE and granted only on request. Abstracting is permitted with mention of source.

Thirty days advance notice is required for change of address. Price per copy: members of the Institute of Radio Engineers, one additional copy $1.25; non-members $2.25. Yearly subscription price: to members $9.00, one additional subscription $13.50; to non-members in United States, Canada, and U. S. Possessions $18.00; to non-members in foreign countries $19.00. Second-class postage paid at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of Febuary 28, 1925, embodied in Paragraph 4, Section 412. P. L. and R., authorized October 26, 1927. Printed in U.S.A.



Miniature

PROKAR® 'D' Molded Capacitors

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Sprague's new and improved PROKAR 'D' Molded Tubular Capacitors meet the need for *ever smaller* molded capacitors capable of withstanding 125 C operation in military, commercial, and industrial electronics.

Key to the new design is an improved processing technique which greatly increases moisture resistance. The new dual dielectric used in Type 150P Capacitors combines the dielectric strength of the highest grade capacitor tissue with the effective moisture resistance of plastic film, giving these miniature units high insulation resistance plus extended life at 125 C. The impregnant used is still the same exclusive high temperature organic material which marked a milestone in molded capacitor development for the original Prokar series.

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Here is the second of two reports on work we have recently completed to help improve our nation's airports. This report, prepared by R. C. Wheeler and K. G. Grossman, describes some of the mathematical considerations.

Analyzing Airport Design

PART II

In the previous AIL monograph, the use of steady-state queuing theory was illustrated with applications to airport operations. However, there are two important reasons why it is advisable to extend the concept to time-dependent queuing theory.

First, an airport operation by nature is characteristically a time-dependent one. For example, at 5:00 A.M. one would not expect to find much traffic, whereas at 8:00 A.M. heavy traffic could be anticipated. In order to apply steady-state concepts we must know how long it takes for steady-state conditions to be achieved—for example, is steady state achieved in 1 hour, 10 hours, or in fact never achieved.

Second, the use of time-dependent queuing models permits an analysis of conditions when airports are overloaded. For overloaded periods, or peaks, steady-state methods offer no help whatsoever.

In this month's article we would like to describe the nature of this work and its applicability to airport analysis. Several mathematical models have potential application to airport analysis. The most realistic one developed thus far requires about seven parameters to define it. To gain insight into the time-dependent behaviour, it was deemed prudent to select a simple model. Consequently, a first-come, first-served model with Poisson input and constant service was selected. In fact, this model does represent quite well some aspects of airport operation.

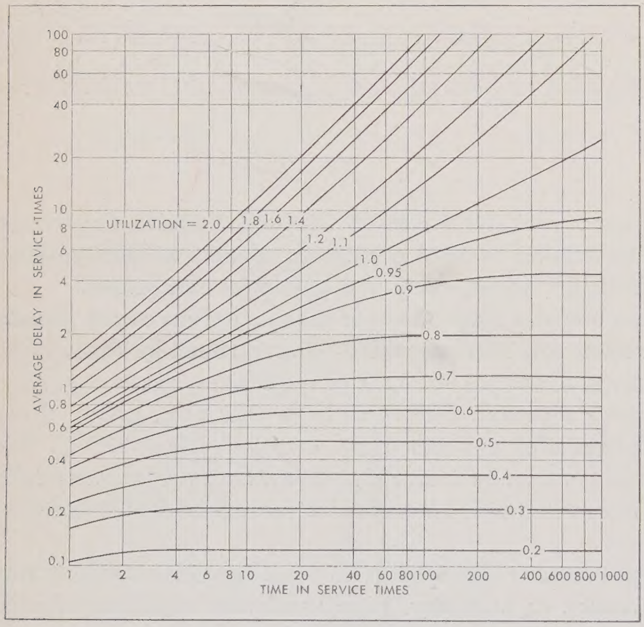

Figure 1. Time-Dependent Average Delay for First-Come, First-Served, Constant Service Queue.

Although this particular model has received considerable attention in the past from a steady-state point of view, relatively little attention has been given to the time-dependent case. Moreover, previous analyses have usually studied the number of units present in the system, whereas this analysis studies the delay directly.

The method used in the solution of this problem employed finite differences. By subdividing the time period for analysis into suitably small intervals of equal size, we are able to obtain the distribution of delay at the end of these intervals. Since the distribution of delay at the end of one interval depends only upon the distribution of delay at the end of the preceding interval, there is a simple relationship between these two distributions. By means of generating functions,

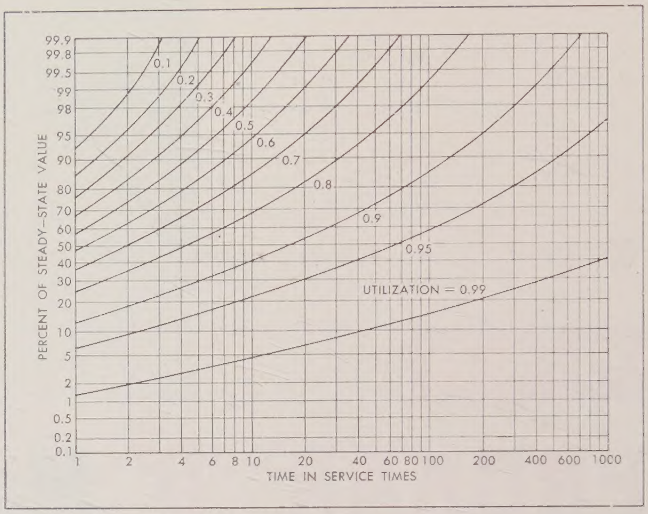

Figure 2. Time-Dependent Average Delay Shown as Fraction of Steady-State Value.

we can express this relationship between these two distributions by the following expressions:

$$g_{i+1}(S) = \frac{1 - e^{-\lambda(1-S)}}{1 - S} + \frac{e^{-\lambda(1-S)}[g_i(S) - g_i(0)]}{S}$$

where

$$g_i(S) = \sum_{k=0}^{\infty} a_{ki} Sk$$

where a_{ki} = probability of delay greater than k increment of time at time increment i

A series of computations were performed for different utilizations where utilization is defined as the product of the demand rate times the average service time. The initial conditions for all computations were that the system was assumed empty. These computations were performed on a 709 electronic computer. Figure 1 shows the relationship of average delay as a function of time (measured in service time) for a variety of utilizations. Note that for utilizations less than 1 the delay curves level out to their steady-state values. However, for utilizations 1 and greater, the average delay continues to rise with time. Figure 2 shows for utilizations less than 1 how fast steady state is achieved in terms of the percent of steady state. For example, for a utilization of 0.99 at 100 service times, the average delay has risen to about 15 percent of the final steady-state value, and even at 1000 service times the average delay has risen to scarcely 40 percent. These curves thus indicate the value of time-dependent concepts when applied to queuing operation.

A complete bound set of our fifth series of articles is available on request. Write to Harold Hechtman at AIL for your set.

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In a moment a new satellite will streak into view. Bell Laboratories may help guide it into orbit, for few are so eminently qualified in the science of missile guidance. Bell Laboratories' Command Guidance System has guided such trailblazers as Tiros and Echo into precise orbits. The same system will guide more new satellites into predetermined orbits as Bell Laboratories continues pioneering in outer space to improve communications on earth.
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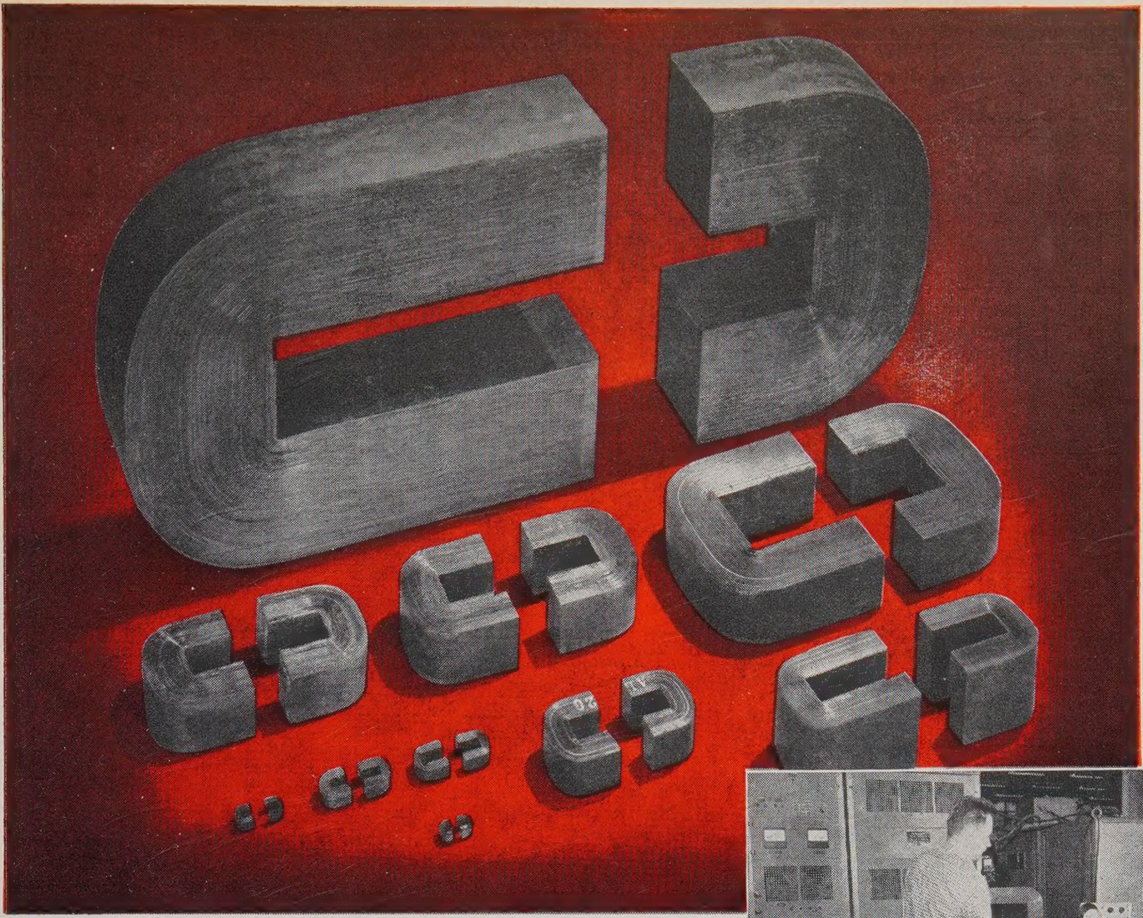
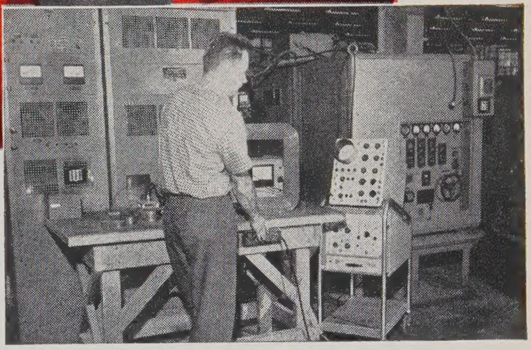


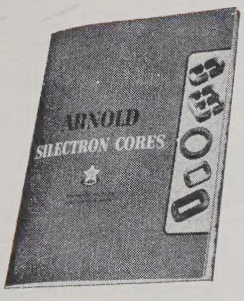


7372D

Arnold Pulse Transformer Cores are individually tested under actual pulse conditions

Here's technical data on

ARNOLD SILECTRON CORES



Bulletin SC-107 A . . . this newly-reprinted 52-page bulletin contains design information on Arnold Tape Cores wound from Silectron (grain-oriented silicon steel). It includes data on cut C and E cores, and uncut toroids and rectangular shapes. Sizes range from a fraction of an ounce to more than a hundred pounds, in standard tape thicknesses of 1, 2, 4 and 12 mils.

Cores are listed in the order of their power-handling capacity, to permit easier selection to fit your requirements, and curves showing the effect of impregnation on core material properties are included. A valuable addition to your engineering files—*write for your copy today.*

ADDRESS DEPT. P-5

The inset photograph above illustrates a special Arnold advantage: a 10-megawatt pulse-testing installation which enables us to test-prove pulse cores to an extent unequalled elsewhere in the industry.

For example, Arnold 1 mil Silectron "C" cores—supplied with a guaranteed minimum pulse permeability of 300—are tested at 0.25 microseconds, 1000 pulses per second, at a peak flux density of 2500 gausses. The 2 mil cores, with a guaranteed minimum pulse permeability of 600, receive standard tests at 2 microseconds, 400 pulses per second, at a peak flux density of 10,000 gausses.

The test equipment has a variable range which may enable us to make special tests duplicating the actual operating conditions of the transformer. The pulser permits tests at .05, .25, 2.0 and 10.0 microsecond pulse duration, at repetition rates varying anywhere from 50 to 1000 pulses per second.

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Meetings with Exhibits

● As a service both to Members and the industry, we will endeavor to record in this column each month those meetings of IRE, its section and professional groups which include exhibits.

Δ

May 8-10, 1961

National Aerospace Electronics Conference (NAECON), Miami & Dayton-Biltmore Hotels, Dayton, Ohio.
Exhibits: Mr. Robert J. Stein, 136 W. Second St., Rm. 202, Dayton 2, Ohio.

May 9-11, 1961

Western Joint Computer Conference, Ambassador Hotel, Los Angeles, Calif.
Exhibits: John H. Whitlock Associates, 253 Waples Mills Road, Oakton, Va.

May 22-24, 1961

Fifth National Global Communications Symposium (GLOBECOM V), Sherman Hotel, Chicago, Ill.
Exhibits: Mr. Fred Hilton, Motorola, Inc., 4501 W. Augusta Blvd., Chicago, Ill.

May 22-24, 1961

National Telemetering Conference, Sheraton Towers Hotel, Chicago, Ill.
Exhibits: Mr. Frank Finch, 795 Gladys Ave., Long Beach 4, Calif.

June 6-8, 1961

Armed Forces Communications & Electronics Show, Sheraton Park and Shoreham Hotels, Washington, D.C.
Exhibits: Mr. William C. Copp, 72 W. 45th St., New York 36, N.Y.

June 13-14, 1961

Fifth National Conference on Product Engineering & Production, Philadelphia, Pa.
Exhibits: Mr. Paul J. Riley, Radio Corp. of America, Building 10-6, Camden 2, N.J.

June 19-20, 1961

Second National Conference on Broadcast and Television Receivers, O'Hare's Inn, Des Plaines, Ill.
Exhibits: Mr. Ray Lee, Philco Corp., 6957 West North Ave., Oak Park, Ill.

June 26-28, 1961

Fifth National Convention on Military Electronics, Shoreham Hotel, Washington, D.C.
Exhibits: Mr. L. David Whitelock, 6514 Greentree Road, Bethesda 14, Md.

July 16-21, 1961

Fourth International Conference on Medical Electronics & Fourteenth Conference on Electrical Techniques in Medicine & Biology, Waldorf-Astoria Hotel, New York, N.Y.
Exhibits: Mr. Lewis Winner, 152 W. 42nd St., New York 36, N.Y.

(Continued on page 10A)

Meetings with Exhibits

(Continued from page 8A)

August 22-25, 1961

Western Electronic Show and Convention (WESCON), Cow Palace and Fairmont Hotel, San Francisco, Calif.
Exhibits: Mr. Don Larson, WESCON, 701 Welch Road, Palo Alto, Calif.

September 6-8, 1961

National Symposium on Space Electronics & Telemetry, Albuquerque, N.M.
Exhibits: Mr. V. V. Myers, 2912 Texas N.E., Albuquerque, N.M.

October 2-4, 1961

Seventh National Communications Symposium, Hotel Utica & Utica Municipal Auditorium, Utica, N.Y.
Exhibits: Mr. R. E. Gaffney, General Electric Co., Light Military Electronics Dept., Utica, N.Y.

October 2-4, 1961

IRE Canadian Convention, Automotive Building, Exhibition Park, Toronto, Canada.
Exhibits: Business Manager, IRE Canadian Convention, 819 Yonge St., Toronto 7, Ontario, Canada.

October 9-11, 1961

National Electronics Conference, Hotel Sherman, Chicago, Ill.
Exhibits: Mr. Rudy Napolitan, National Electronics Conference, 228 N. LaSalle St., Chicago, Ill.

October 23-25, 1961

East Coast Conference on Aeronautical & Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md.
Exhibits: Mr. Robert J. Henderson, Martin Company, Ground Support Equipment Dept., Baltimore, Md.

October 23-26, 1961

Eighth Annual Meeting, Professional Group on Nuclear Science (Aero-Space Nuclear Propulsion), Hotel Riviera, Las Vegas, Nevada
Exhibits: Mr. D. J. Knowles, Union Carbide Company, Oak Ridge National Lab., Oak Ridge, Tenn.

October 26-27, 1961

Electronic Techniques in Medicine & Biology Conference, University of Nebraska, Omaha, Neb.
Exhibits: Mr. Harold G. Beenken, University of Nebraska, College of Medicine, 428 Dewey Avenue, Omaha, Neb.

Δ

Note on Professional Group Meetings: Some of the Professional Groups conduct meetings at which there are exhibits. Working committeemen on these groups are asked to send advance data to this column for publicity information. You may address these notices to the Advertising Department and of course listings are free to IRE Professional Groups.

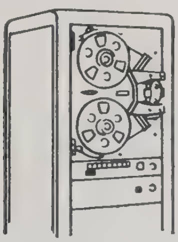
A WORD OF WARNING ABOUT THE NEW ALLUREMENTS OF RECOMP II [and a modest word about price]

Could you be enticed by a computer? Surprisingly, there *are* businessmen and scientists who have allowed their emotions to get quite out of control regarding Recomp II.

And now there is more reason than ever for becoming enamored with this amazing computer. *Three* reasons, to be exact, and all of them new. Hence, our warning to you.

The first reason is, in itself, enough to steal your heart away: it is Recomp II's new reduced lease price. Always the darling of the medium-scale computer user, Recomp II has been so well accepted that it can now be offered at significantly lower terms. And it *still* provides the identical quality, solid-state performance, and features that can't be found on computers costing three times what Recomp II *used* to cost.

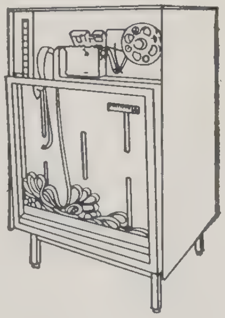
This is heady stuff—but even more enticements lie in wait. You can now add an optional modification to your Recomp II to enlarge its capacity by using magnetic tape. Here you see the new Recomp Magnetic Tape Transport unit.



Naturally it's superbly designed, solid state throughout. But don't let its quietly well-bred air fool you; it has a memory that would stagger an elephant—over 600,000 words. And up to eight of the Transport units can be connected to Recomp II, giving you a computer with a total memory capacity of over 5,000,000 words. Steady there, Mr. Simpson!

The speed of this new magnetic tape control is something to applaud, too: read and write speed is 1850 characters a second; bidirectional search speed is 55 inches per second. Do you begin to see why we warned you about these new allurements of Recomp?

Below you see another new optional feature for your Recomp II: the Facitape tape punch and reader console. It punches 150 characters a second, reads 600 characters a second, and stops on a character. It adjusts to read and punch from 5



through 8 channels. It is versatile, accurate, fast, simple-to-operate, economical, reliable. And it has perfect manners: the mechanical components are completely enclosed in a soundproof housing.

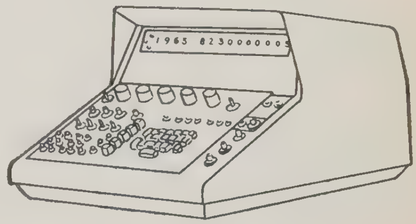
But lest we harp too much on the *new* features of Recomp II, perhaps we had better remind you of some of the extraordinary features that Recomp II *already had*. Features that have always made it the finest computer in the low-priced field.

1] Recomp II is the *only* compact computer with built-in floating point arithmetic. It defies being hemmed in on a problem. With its large capacity it obviates computer-claustrophobia.

2] Recomp II was the first solid-state computer on the market. As you can see by the new features above, Recomp II's scrupulous engineers have seen to it that it remains the *finest* solid-state computer on the market.

3] Recomp II seems to have more built-in features than a dream home kitchen. It has built-in square root command. Built-in automatic conversion from decimal to binary.

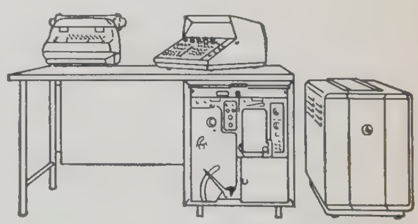
Here you see Recomp II's distinctive keyboard. It looks easy enough to operate—and it *is!* And because Recomp II



requires no specialized talents, anyone with computer problems can be taught to use it.

One look at Recomp II leaves little wonder that even practical people have allowed their hearts to influence them in choosing Recomp II. Without being showy, it is an object of beauty that reflects its supreme precision of performance. Its distinguished exterior bespeaks the ultimate of excellence; *c'est sans pareil.*

But if you want to avoid being captivated by a computer you should know how strong your emotions will run. May we suggest a *test?* Expose yourself to Recomp II. See it in action. Touch it. Feed problems into it. This is the only way to know how you will react to this extraordinary computer. Make a date to see Recomp II right away.



Write AUTONETICS INDUSTRIAL PRODUCTS, Dept. O58, 3400 E. 70th St., Long Beach, Calif.
The Autonetics Division of North American Aviation, Inc.

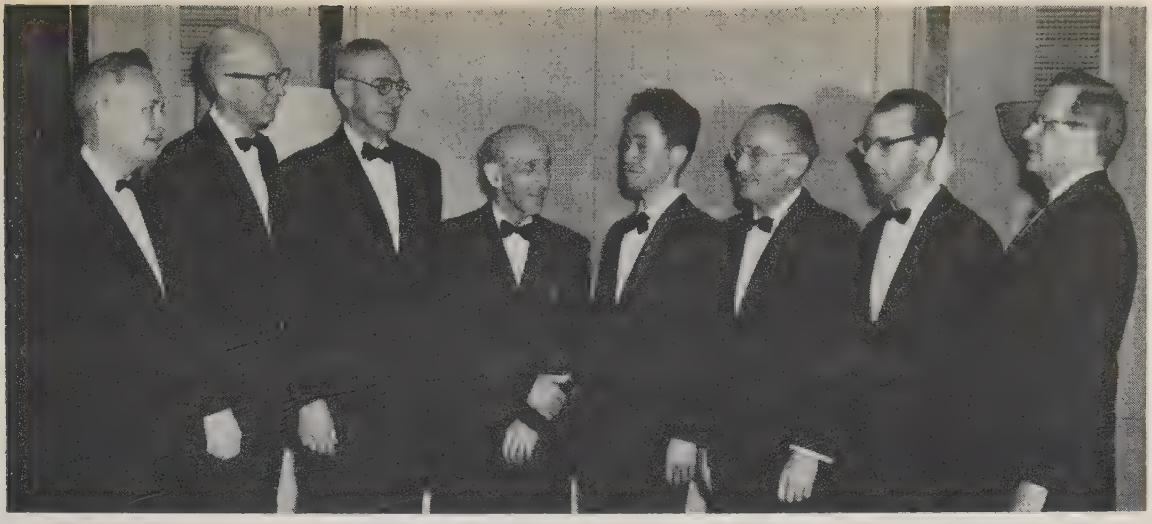
1961 INTERNATIONAL IRE CONVENTION

The above is a view of some of the exhibits on one of the four floors of the Coliseum. There were over 850 exhibitors at this year's IRE Show and more than 25,000 items were displayed in $2\frac{1}{4}$ miles of booths. The final attendance was 67,419.

(*Left*) John F. Byrne, North American Vice President of the IRE; Lloyd V. Berkner, President; George W. Bailey, Executive Secretary and Chairman of the 1961 International Convention; and Franz Ollendorff, Overseas Vice President, at the opening press conference of the Convention.

(*Right*) President L. V. Berkner presents the Professional Group on Bio-Medical Electronics Prize Award in Memory of William J. Morlock to Britton Chance. The award was presented for the first time at the 1961 Annual Banquet.

(*Left*) Dr. Ralph Bown is presented with the 1961 Founders' Award by Dr. Berkner.



(*Left to right*): P. E. Haggerty, principal speaker at the Annual Banquet; C. R. Wischmeyer, spokesman for the newly-elected Fellows; Dr. Ralph Bown, winner of the IRE Founders' Award; Dr. Ernst A. Guillemin, winner of the IRE Medal of Honor; Dr. Leo Esaki, recipient of the Morris N. Liebmann Memorial Prize; H. L. Brueckmann, winner of the Harry Diamond Memorial Prize; Manfred Clynes, winner of the W. R. G. Baker Award; and P. C. Goldmark, the recipient of the Vladimir K. Zworykin Prize. Dr. Esaki accepted the Browder J. Thompson Memorial Prize for Eiichi Goto, who was unable to attend the Convention.

Daniel E. Noble, toastmaster; Patrick E. Haggerty, principal speaker; and Carl R. Wischmeyer, spokesman for the newly-elected Fellows, at the Annual Banquet.

(*Left*) IRE President Lloyd V. Berkner presents the IRE Medal of Honor to Dr. Ernst A. Guillemin (*right*), at the Annual Banquet.

The Annual IRE Banquet, a view of which is shown above, was held in the Grand Ballroom of the Waldorf-Astoria Hotel.

Current IRE Statistics

(As of March 31, 1961)

Membership—89,283
Sections*—110
Subsections*—30
Professional Groups*—28
Professional Group Chapters—286
Student Branches†—204

* See this issue for a list.
† See October, 1960 issue for a list.

Calendar of Coming Events and Authors' Deadlines*

1961

Workshop in Graph Theory, University of Illinois, Urbana, May 6.

5th Midwest Symp. on Circuit Theory, Allerton Park & Urbana Campus, Univ. of Ill., Urbana, May 8–9.

NAECON, Miami & Biltmore Hotels, Dayton O., May 8–10.

Western Joint Computer Conf., Ambassador Hotel, Los Angeles, Calif., May 9–11.

Microwave Theory and Tech. Nat'l. Symp., Sheraton Park Hotel, Washington, D. C., May 15–17.

GLOBECOM V, Sherman Hotel, Chicago, Ill., May 22–24.

Nat'l. Telemetering Conf., Sheraton Hotel, Chicago, Ill., May 22–24.

Electro-Optical Devices Symp., Los Angeles, Calif., May.

3rd Nat'l. Symp. on Radio Frequency Interference, Sheraton-Park Hotel, Washington, D. C., June 12–13.

5th Nat'l. Symp. on Product Engrg. and Production, Hotel Sheraton, Philadelphia, Pa., June 14–15.

Chicago Spring Conf. on Broadcast and Television Receivers, O'Hare Inn, Des Plaines, Ill., June 19–20.

MIL-E-CON 1961, Shoreham Hotel, Washington, D. C., June 26–28.

JACC, Univ. of Colorado, Boulder, June 28–30.

Int'l. Conf. on Elec. Engrg. Education, Syracuse, N. Y., June.

4th Int'l. Conf. On Medical Electronics & 14th Conf. on Elec. Techniques in Medicine & Biology, Waldorf-Astoria Hotel, N. Y., N. Y., July 16–21. (DL*: April 1, 1961, H. P. Schwan, Moore School of E.E., Philadelphia 4, Pa.)

WESCON, San Francisco, Calif., Aug. 22–25. (DL*: May 1, E. W. Herold, WESCON North Calif. Office, 701 Welch Rd., Palo Alto, Calif.)

3rd Int'l. Conf. on Analog Computation, Belgrade, Sept. 4–9.

* DL = Deadline for submitting abstracts.

(*Continued on page 15A*)

Gordon Research Conferences to Include July Session on Semiconductors

The 1961 Gordon Research Conference on Chemistry and Metallurgy of Semiconductors will be held at Tilton School, Tilton, N. H., from July 10–14. A. J. Rosenberg and P. Egli are Chairman and Vice Chairman, respectively, of the one-week conference. The program sessions will comprise the following general topics: epitaxial crystal growth, crystal growth at high pressures, chemical bonds and electron energy bonds, organic conductors, and electroluminescence.

The Gordon Research Conferences for 1961 will be held from June 12 to September 1 at Colby Junior College, New London, N. H., New Hampton School, New Hampton, N. H., Kimball Union Academy, Meriden, N. H., as well as at the Tilton School.

The series of conferences was established to stimulate research in universities, research foundations, and industrial laboratories. This purpose is achieved by an informal type of meeting consisting of scheduled lectures and discussion groups.

The purpose of the program is to bring experts up to date on the latest developments, to analyze the significance of these developments and to provoke suggestions concerning the underlying theories and profitable methods of approach for making progress. The review of known information is not desired.

In order to protect individual rights and to promote discussion, it is an established requirement of each conference that no information presented is to be used without specific authorization of the individual making the contribution, whether in formal presentation or in discussion. Scientific publications are not prepared as emanating from the conferences.

Attendance at the Conferences is by application. Individuals interested in attending are requested to send their applications to the Director at least two months prior to the date of the Conference. All applications must be submitted in, duplicate on the standard application form which may be obtained by writing to the office of the Director.

Requests for attendance at the Conferences, or for additional information, should be addressed to W. George Parks, Director, Dept. of Chemistry, University of Rhode Island, Kingston, R. I. From June 12 to September 1, 1961, mail for the office of the Director should be addressed to Colby Junior College, New London, N. H.

AIEE Announces 1962 Winter Meeting

The American Institute of Electrical Engineers now issues a call for papers for its 1962 Winter General Meeting, which will be held in New York, N. Y., on January 28–February 2, 1962. Prospective authors are invited to submit papers on "Kilomegacycle Computing Systems" for presentation at sessions during this meeting, and for publication as AIEE Transaction Papers. Papers related to this topic will be kept together to enhance their value.

July 1, 1961 is the deadline for 100-word abstracts and 500-word informal summaries (or the entire paper, if available). Correspondence regarding these sessions, as well as requests for further information, should be directed to J. H. Wright (Papers Chairman), Div. 12, U. S. National Bureau of Standards, Washington 25, D. C.

(*Left to right*): Walter McClelland, Vice Chairman of the Chicago Section; Ray DeCola, Secretary-Treasurer of the National Administrative Committee of the PGBTR; John Bell, member of the National Administrative Committee of the PGBTR; and Jack Bridges, Chairman of the Chicago Spring Conference on Broadcast and Television Receivers, participated in a joint briefing session during which full support was pledged for the Conference, which will be held in DesPlaines, Ill., on June 19–20, 1961.

TRANSACTIONS Binders Now Available

Four-inch maroon binders of Spanish-grain fabrikoid, equipped with 24 steel blades, are now available for storing issues of the IRE TRANSACTIONS.

The binders, with the words "IRE Professional Group TRANSACTIONS" imprinted in gold, may be ordered at $3.00 each from IRE Headquarters, 1 East 79th St., New York 21, N. Y. Any or all of the following three items can also be imprinted at an additional charge of 50¢ per item: your name; Group name; year. Also, extra blades for the binders may be purchased at 10¢ each.

CALL FOR PAPERS ISSUED BY CANADIAN CONFERENCE

An invitation for the submission of papers for presentation at the technical sessions of the IRE Canadian Electronics Conference has been issued. The Conference is to be held at Exhibition Park in Toronto, October 2–4, 1961.

In issuing the invitation, A. R. Low, Chairman of the Conference's Technical Program Committee, stressed that it was not necessary for an author to be a member of the Institute of Radio Engineers.

The paper may be on any topic of general interest to engineering, management, government or college members of the IRE; the topic may deal with any aspect of the electronics field, such as basic research, design, techniques, production, reliability and education.

Authors are asked to submit a 500- to 1000-word summary of the proposed paper. Authors of selected papers will be asked to supply a 100-word abstract for inclusion in the conference program pamphlet, and a biographical sketch.

The publication of a digest of the papers presented is being considered. This would be available at Conference registration. For such a digest, authors would be asked to submit diagrams along with a 500- to 1000-word summary of the paper in a form suitable for direct photographic reproduction.

Prospective authors should contact: A. R. Low, IRE Canadian Electronics Conference Headquarters, 1819 Yonge St., Toronto, Ont., Canada.

NEC ISSUES CALL FOR PAPERS

The Seventeenth Annual National Electronics Conference, which will be held at the International Amphitheatre, Chicago, Ill., on October 9–11, 1961, now issues a call for papers for presentation at the meeting.

The papers program of the NEC covers the entire range of technology of interest to electronic engineers—components, devices, design, techniques, applications, management, etc. The following list is indicative of the scope of the Conference, but should not be considered all-inclusive:

- Antennas
- Modern Microwave Techniques
- Millimeter Wave Techniques
- Optical Communications
- Audio and Ultrasonic Engineering
- Magnetohydrodynamics
- Microelectronics
- Solid State Devices and Circuits
- Transistor Circuit Application
- Parametric Devices and Techniques
- Space Communications
- Man in Space
- Logic and Switching Theory
- Digital Computer Design
- Digital Computer Techniques
- Digital System Components
- Digital Data Transmission
- Data Processing
- Engineering Management
- Learning and Adaptive Systems
- Engineering Cryogenics
- Network Theory
- Amplifier Circuit Theory
- Instrumentation

Two $500 awards will be given for the best papers submitted at the Conference. Last year's winners were L. P. Huelsman of the University of Arizona, Tucson, for the best over-all paper, entitled, "Active RC Synthesis With Prescribed Sensitivity," and Dr. R. M. Lerner, of the Massachusetts Institute of Technology, Cambridge, for the best tutorial paper, "Modulation and Signal Selection from Digital Data Systems."

Authors whose papers are selected for presentation at the Conference will have their papers included in the *Proceedings*, receive complimentary advance registration, a complimentary copy of the *Proceedings*, and an author's ribbon badge.

For further information, contact: W. L. Firestone, Motorola, Inc., 4501 W. Augusta Blvd., Chicago 51, Ill. Prospective authors are advised to do so as soon as possible. The final copies of the completed papers must be submitted by September 6, 1961.

CHICAGO SPRING CONFERENCE ANNOUNCES JUNE MEETING

The Chicago Spring Conference on Broadcast and Television Receivers is to be held on June 19–20, 1961 at the O'Hare Inn, Des-Plaines, Ill. The Conference is sponsored by the Professional Group on Broadcast and Television Receivers and the Chicago Section of the IRE. Twenty papers are scheduled for delivery during four technical sessions. Topics will cover transistorized radio and television circuits, satellite broadcasting, FM stereo broadcasting and receivers, subscription television, and application of solid state devices to entertainment products.

Exhibits of particular interest to the entertainment field will be on display in space adjacent to the technical sessions.

A block of rooms at the O'Hare Inn is available for reservation by attendees of the Conference.

Registration fees for nonmembers of the IRE will be $4.00; the fee for IRE members $2.00; and lunches $2.00. A luncheon speaker will deliver a welcoming address on Monday.

For further information on registration, contact; Chad Pierce, Wells-Gardner, 2701 N. Kildare Ave., Chicago, Ill.

Calendar of Coming Events and Authors' Deadlines*

(*Continued from page 14A*)

1961 Nat'l. Symp. on Space Electronics and Telemetry, Albuquerque, N. M., Sept. 6–8.

Joint Nuclear Instrumentation Symp., North Carolina State College, Raleigh, N. C., Sept. 6–8.

Engrg. Writing and Speech Symp., Bellevue Stratford Hotel, Philadelphia, Pa., Sept. 14–15.

9th Ann. Engrg. Management Conf., New York, N. Y., Sept. 14–16.

10th Ann. Industrial Electronics Symp., Boston, Mass. Sept. 20–21.

7th Nat'l. Communications Symp., Utica, N. Y., Oct. 2–4. (DL*: June 1, R. K. Walker, 34 Bolton Rd., New Hartford, N. Y.)

IRE Canadian Electronics Conf., Automotive Bldg., Exhibition Park, Toronto, Canada, Oct. 2–4.

Nat'l. Electronics Conf., Int'l. Amphitheatre, Chicago, Ill., Oct. 9–11.

5th Nat'l. Symp. on Engrg. Writing and Speech, Kellogg Ctr. for Continuing Education, Michigan State Univ., East Lansing, Oct. 16–17. (DL*: July 15, J. Chapline, Philco Corp., Computer Div., 3900 Welsh Rd., Willow Grove, Pa.)

East Coast Conf. on Aerospace & Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md., Oct. 23–25.

URSI-IRE Fall Mtg., Univ. of Texas, Austin, Oct. 23–25.

PGNS 8th Ann. Mtg., Hotel Riviere, Las Vegas, Nev., Oct. 23–26.

Int'l. Symp. on Aero-Space Nuclear Propulsion, Las Vegas, Nev., Oct. 23–26. (DL*: July 1, P. M. Uthe, Univ. of Calif., Lawrence Radiation Lab., Box 808, Livermore, Calif.)

Symp. on Instrumentation Facilities for Biomedical Res., Sheraton Fontenelle Hotel, Omaha, Neb., Oct. 26–27.

1961 Electron Devices Mtg., Sheraton-Park Hotel, Washington, D. C., Oct. 26–28.

Radio Fall Mtg., Hotel Syracuse, Syracuse, N. Y., Oct. 30–31.

6th Ann. Special Technical Conf. on Nonlinear Magnetics, Statler Hilton Hotel, Los Angeles, Calif., Nov. 6–8. (DL*: June 1, T. Bernstein, Space Technology Labs., Inc., P. O. Box 95001, Los Angeles, Calif.

7th Ann. Conf. on Magnetism and Magnetic Materials, Hotel Westward Ho, Phoenix, Ariz., Nov. 13–16.

NEREM, Boston Mass., Nov. 14–16.

MAECON, Kansas City, Mo., Nov. 14–16.

PGVC Conf., Hotel Leamington, Minneapolis, Minn., Nov. 30–Dec. 1.

Eastern, Joint Computer Conf., Sheraton-Park Hotel, Washington, D. C., Dec. 12–14.

1962

8th Nat'l. Symp. on Reliabi ity and Quality Control, Statler Hilton Hotel, Washington, D. C., Jan 9–11. (DL*: May 15, 1961, E. F. Jahr, IBM Corp., Owego, N. Y.)

* DL = Deadline for submitting abstracts.

1961 NTC Announces Program Sessions

The theme of the 1961 National Telemetering Conference is "Science and Education in Telemetry." It will be held May 22–24, 1961, at the Sheraton Chicago Hotel, Chicago, Ill. The IRE, AIEE, ARS, IAS, and ISA, are sponsoring the Conference. Technical program sessions and chairmen have been announced as follows:

Monday Morning, May 22

Workshop: Telemetry Standards, *A. E. Bentz, Sandia Corp.*

Industrial Telemetering, *W. E. Phillips, Leeds and Northrup Co.*

Monday Afternoon

Transducers, *R. G. Jewell, General Electric Co.*

Advanced Systems Techniques I, *W. Van Dyke, Douglas Aircraft Co.*

Data Processing and Presentation, *T. J. Martin, Boeing Airplane Co.*

Tuesday Morning, May 23

Workshop: Telemetry in Europe, *A. P. Gruer, Sandia Corp.*

Signal Conditioning, *R. W. Sanders, Space Electronics.*

Flight Test Data Systems, *J. Mallernee, North American Aviation.*

Tuesday Afternoon

Workshop: Telemetry Education, *W. J. Mayo-Wells, George Washington University.*

Advanced Systems Techniques, *A. C. Cole, The Martin Co.*

Wednesday Morning, May 24

PCM Systems I, *M. G. Pawley, United States Naval Ordnance Lab.*

Underwater Measurement, *M. A. Lowy, Consultant.*

Wednesday Afternoon

Bio-Medical Telemetering, *M. McLennan, Wright-Patterson Air Force Base.*

PCM Systems II, *F. Karabaich, Wright-Patterson Air Force Base.*

RF Components and Techniques, *T. Eccles, Microdot, Inc.*

Call for Papers For Las Vegas Symposium

An International Symposium on Aerospace Nuclear Propulsion will be held in Las Vegas, Nev., October 23–26, 1961. This symposium is sponsored jointly by the PGNS, the AEC and the NASA. The meeting will be unclassified and will include, among other topics, a review of the Rover, Pluto, and ANP programs.

Papers are being requested for presentation at the meeting. Subject matter includes: instrumentation, control systems, engine simulation, and engine dynamics aspects of aerospace nuclear propulsion and nuclear auxiliary power fields, as well as radiation and temperature effects on instruments and controls.

A rough draft of each paper and a 500-word abstract should be submitted by July 1, 1961, to P. M. Uthe, University of California, Lawrence Radiation Lab., Box 808, Livermore, Calif. The length of each paper should be approximately 25 minutes presentation time.

MISCELLANEOUS IRE PUBLICATIONS AVAILABLE

The following issues of Miscellaneous Publications are available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y., at the prices indicated below:

Meetings	Publications	Price per Copy
Aeronautical and Navigational Electronics Conference	*Proceedings of the 5th Annual East Coast ANE Conference,* held October 27–28, 1958 in Baltimore, Md.	$5.00*
Electronic Computer Conferences	*Proceedings of the Joint AIEE IRE ACM Eastern Conference,* held December 10–12, 1951 in Philadelphia, Pa.	3.50
	Proceedings of the Joint AIEE IRE ACM Eastern Conference, held December 8–10, 1954 in Philadelphia, Pa.	3.00
	Proceedings of the Joint AIEE IRE ACM Eastern Conference, held November 7–9, 1955 in Boston, Mass.	3.00
	Proceedings of the Joint AIEE IRE ACM Eastern Conference, held December 13–15, 1960 in New York, N. Y.	3.00
	Proceedings of the Joint AIEE IRE ACM Western Conference, held March 1–3, 1955 in Los Angeles, Calif.	3.00
	Proceedings of the Joint AIEE IRE ACM Western Conference, held February 7–9, 1956 in San Francisco, Calif.	3.00
	Proceedings of the Joint AIEE IRE ACM Western Conference, held May 3–5, 1960 in San Francisco, Calif.	3.00
Magnetic Amplifiers Conference	*Proceedings of the Conference on Magnetic Amplifiers,* held April 5–6, 1956 in Syracuse, N. Y.	4.00
Bio-Medical Electronics Bibliographies	*Bibliography on Medical Electronics,* June, 1958 issue	2.50
	Bibliography on Medical Electronics, June, 1959 issue, (Supplement I)	2.50
	Bibliography on Medical Electronics, July, 1960 issue, (Supplement II)	2.50
Military Electronics Conferences	*Proceedings of the 1st National Convention,* held June 17–19, 1957 in Washington, D. C.	5.00†
	Proceedings of the 2nd National Convention, held June 16–18, 1958 in Washington, D. C.	5.00†
	Proceedings of the 3rd National Convention, held June 29–July 1, 1959 in Washington, D. C.	4.00
	Proceedings of the 4th National Convention, held June 27–29, 1960 in Washington, D. C.	5.00
Radio Frequency Interference Symposium	*Digest of the 2nd National Symposium,* held June 13–14, 1960 in Washington, D. C.	2.00
Reliability and Quality Control in Electronics Symposia	*Proceedings of the 1st National Symposium,* held November 12–13, 1954 in New York, N. Y.	5.00
	Proceedings of the 2nd National Symposium, held January 9–10, 1956 in Washington, D. C.	5.00
	Proceedings of the 3rd National Symposium, held January 14–16, 1957 in Washington, D. C.	5.00
	Proceedings of the 4th National Symposium, held January 6–8, 1958 in Washington, D. C.	5.00

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* IRE member rate—$3.50
† IRE member rate—$3.00

MISCELLANEOUS IRE PUBLICATIONS AVAILABLE

The following issues of Miscellaneous Publications are available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, New York, at the prices indicated below:

Meetings	Publications	Price per Copy
	Proceedings of the 5th National Symposium, held January 12–14, 1959 in Philadelphia, Pa.	$5.00
	Proceedings of the 6th National Symposium, held January 11–13, 1960 in Washington, D. C.	5.00
	Proceedings of the 7th National Symposium, held January 9–11, 1961 in Philadelphia, Pa.	5.00
Space Electronics and Telemetry Symposium, plus Telemetering Conference	*Proceedings of the 1959 National SET Symposium,* held September 28–30, 1959 in San Francisco, Calif.	5.00
	Proceedings of the 1960 National Telemetering Conference, held May 23–25, 1960 in Santa Monica, Calif.	6.50‡

‡ IRE member rate—$4.50

Conference on Nonlinear Magnetics Scheduled for November

The Sixth Annual Special Technical Conference on Nonlinear Magnetics will be held at the Statler Hilton Hotel in Los Angeles, Calif., on November 6–8, 1961. This conference is under the auspices of the AIEE and the IRE. The theme will be "Magnetics in the Space Age."

Papers which will deal with nonlinear magnetics, magnetic amplifiers, computer applications, combined semiconductors and nonlinear magnetics are now being solicited. Final manuscripts are due on June 1, 1961. For further information, p ease contact: Dr. T. Bernstein, Space Technology Labs., Inc., P. O. Box 95001, Los Angeles 45, Calif.

JTAC Pays Tribute To Dr. Hogan

The Joint Technical Advisory Committee passed the following resolution at its meeting on January 19, 1961, as a tribute to the late Dr. John V. L. Hogan:

"The Joint Technical Advisory Committee and its sponsors (The Institute of Radio Engineers and the Electronic Industries Association) wish to express their deep sense of loss on the death of Dr. John V. L. Hogan, a former Chairman and active member of JTAC since its inception in July, 1948.

"Dr. Hogan's balanced judgment and inherent kindness toward his fellow man contributed greatly toward JTAC's sound and impartial decisions. Through his unselfish devotion he served well and long the profession and the electronics industry. In addition to his contributions to the JTAC, Dr. Hogan was truly one of the few pioneers of the field known today as electronics. He was one of the three founders of IRE in 1912.

"Dr. Hogan is sorely missed by JTAC and the many other activities where his cheerful counsel, keen intellect and broad standing contributed so largely to our technical progress, national security and welfare.

Air Force MARS Announces Schedule

The schedule of broadcasts of the Air Force MARS Eastern Technical Net, operating Sundays from 2 to 4 P.M. EDT, at 3295, 7540, and 15,715 kc, has been announced as follows:

May 14—"Semiconductors," Dr. A. I. Bennett, Advisory Physicist, Westinghouse Research Laboratories.

May 21, 28, June 4, and 11—"Review of Basic Physics," staff engineers of the General Electric Company.

PGRFI to Hold June Conference

The Third National Symposium on Radio Frequency Interference will be held in Washington, D. C., at the Sheraton Park Hotel on June 12–13, 1961. The Chairman of the Symposium is D. R. J. White, Vice President and Director of Research at Frederick Research Corporation, West Wheaton, Md. The symposium program will represent a departure from past symposia in that it will make strong use of panels to discuss highly controversial matters. Emphasis will be placed on the DOD Tri-Service Radio Frequency compatibility program. For further information, contact: E. F. Mischler, Chairman of Public Relations at National Engineering Service, Washington, D. C., EXecutive 3-7065.

EWS Symposium Issues Call for Papers

A call for papers has been issued by the East Lansing Symposium on Engineering Writing and Speech, which will be held on October 16–17, 1961, at the Kellogg Center for Continuing Education, Michigan State University, East Lansing.

The papers should deal with the theme "Communicating Ideas—The Modern Engineer's Function." Prospective authors are requested to send 500-word summaries by July 15, 1961 to: J. D. Chapline, Philco Corp., 3900 Welsh Road, Willow Grove, Pa.

Soviet Physics-Doklady To Appear Monthly

The American Institute of Physics has announced that its *Soviet Physics-Doklady,* which offers brief reports on Russian research encompassing the entire spectrum of pure and applied physics, will now appear monthly, starting with the July, 1961 (Vol. 7, No. 1) issue. Previously the journal had been published every other month.

Soviet Physics-Doklady is a translation of the physics sections of *Doklady Akademii Nauk SSSR*—the Proceedings of the Academy of Sciences, USSR. Inquiries regarding the journal should be sent to the American Institute of Physics, 335 E. 45 St., New York 17, N. Y.

Globecom V Meets in Chicago

Arrangements have been completed for the Fifth National Symposium on Global Communications, which will convene at the Hotel Sherman in Chicago, Ill., on May 22–24, 1961. The conference is being sponsored by the AIEE and the IRE Professional Group on Communications Systems. According to Dr. W. L. Firestone, General Chairman of GLOBECOM V, over one thousand engineers, scientists, and managers are expected to participate in the Symposium. Eighteen sessions of technical papers, covering all phases of the communications art, will be presented. Space communications, switching and transmission systems, and data handling will be featured under the theme "Communications on a Global Scope." The central social function will be the Symposium luncheon on Tuesday, May 23. Dr. G. E. Mueller, Vice President, Space Technology Laboratories, Los Angeles, Calif., will be the principal speaker. In addition to the technical papers program, some two dozen manufacturing and engineering organizations will be displaying the latest in communications equipment, processes and technical advances.

A Convention Record, containing abbreviated versions of each paper, will be provided to all participants as they register at the conference. This will aid the speakers and audience in maintaining the high professional settings for the Symposium.

(*Continued on page 20A*)

Six *Styroflex®* *Coaxial Cables* on One Tower

...feed antennas for WIIC (TV) and WWSW-FM in Pittsburgh!

A whole "family" of six Styroflex® coaxial cables is currently in use on a single 812-foot tower in Pittsburgh, Pa. The six cable runs act as transmission lines for television station WIIC and for radio station WWSW-FM.

Two 6⅛" Styroflex® cables serve as main transmission lines terminating in the main antenna. One of these is operating at a transmitter output of 95 KW Peak Visual Power; the other handles a transmitter output of 47.5 KW cw Aural Power. A pair of 3⅛" cables are used to connect the 11 KW auxiliary transmitter to separate auxiliary antennas. Another 3⅛" cable is used as the primary antenna feed for WWSW-FM, with a 1⅝" cable acting as a standby line for this station.

Long operating life and stable electrical properties in a wide variety of climatic conditions make Styroflex® cable ideal for many kinds of applications. Manufactured in 1000-foot lengths, Styroflex® cable eliminates the need for 20-foot interval connectors that can cause gas leakage problems in rigid lines. The longer cable lengths also simplify installation of cable runs.

The performance record of Styroflex® cable has earned for it an outstanding reputation in telemetry, guided missile and mass communication applications. If you have need for a superior high frequency cable with proven properties, we suggest you consider Styroflex®.

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PD

Moderators for the technical sessions have been announced by Donald Campbell, Program Chairman. They are:

R. D. Campbell, AT&T, **Space Communication I.**

R. E. Stoffels, Automatic Electric Labs., **Data Transmission.**

A. A. MacDonald, Motorola, **Communications Systems.**

W. K. MacAdam, Vice President, AT&T, **A Large Scale Four-Wire Switched Communications Network for Military Communications.**

J. N. Petrie, Automatic Electric, **System Performance.**

W. Lyons, RCA, **HF Communications.**

Prof. E. C. Jordan, University of Illinois, **Space Communication II.**

Col. J. Z. Millar, Western Union, **Modulation Techniques.**

Col. W. D. Joslin, U. S. Army Signal Corps, **Pacific Scatter Communication System.**

H. F. May, Bell Telephone Labs., **Switching Systems I.**

Paul Hurtel, Jr., Collins Radio Co., **Microwave Radio Relay.**

Dr. H. P. Messinger, ITT Kellogg, **Reliability.**

Robert L. Vader, Lockheed, **Space Communication III.**

V. N. Vaughan, AT&T, **Data Handling.**

Nelson B. Tharp, Westinghouse, **Forward Scatter.**

R. D. Slayton, Teletype, **Switching Systems II.**

Prof. A. H. Rubenstein, Northwestern University, **System Planning.**

R. C. Benoit, Jr., USAF, RADC, **Speech Compression and Spectrum Sharing.**

The Symposium Committee is composed of:

General Chairman: Dr. W. L. Firestone, Motorola, Inc.

Vice Chairman (AIEE): S. R. Collis, Illinois Bell Telephone Co.

Vice Chairman (IRE): R. H. Maier, ITT Kellogg.

Executive Secretary: C. F. Wittkop, Motorola, Inc.

Technical Program Chairman: D. C. Campbell, ITT Kellogg.

Arrangements Chairman: R. J. Hajek, Commtronix Engineering.

Public Relations and Publicity: R. D. Slayton, Teletype Corporation.

Publications Chairman: E. F. Dirsa, Admiral Corporation.

Finance Chairman: R. E. Morrison, Illinois Bell Telephone Co.

Exhibits Chairman: F. L. Hilton, Motorola, Inc.

An advance registration drive opened in late March. Further details on the program and registration may be obtained from the Executive Secretary, C. F. Wittkop, Motorola, Inc., 1450 N. Cicero Ave., Chicago 51, Ill.

Fifth Annual PGPEP Conference

HOTEL SHERATON, PHILADELPHIA, PA., JUNE 14–15, 1961

P. J. Riley, General Chairman, has announced the technical program and other highlights of the Fifth Annual Conference of the Professional Group on Product Engineering and Production (PGPEP) of the IRE, which will be held in Philadelphia, Pa., June 14–15, 1961, at the Hotel Sheraton. Besides the four technical sessions, the activities will include a luncheon speech by T. A. Smith, Executive Vice President, Industrial Electronic Products Division of RCA; presentation of a science award to an outstanding high school student selected through the science fair program in the Delaware Valley Public Schools; and a speech by a prominent figure in the missile industry. The social activities will also include a choice of three tours for the wives: an historical tour, a tour of department stores, or a tour of a TV station.

Exhibits chairman, Arthur Ansley reports that twenty-five companies are expected to set up exhibits. Exhibitors may contact Mr. Ansley at Ansley Manufacturing Co., New Hope, Pa., for space.

A pre-conference package is available for IRE members for $10.00, which includes registration, a copy of the *Proceedings*, and luncheon. Separate fees are: registration—$3.50 for IRE members and $5.00 for nonmembers; *Proceedings*—$5.00; and luncheon—$4.00. Those interested in attending the conference should contact: W. J. Welsh, Registration Chairman, RCA, Moorestown, N. J., for reservation and registration.

The technical program has been announced as follows:

Wednesday Morning, June 14

Computer and Data Processing

Moderator: *I. Auerbach, Auerbach Electronics.*

"Digital Microelectronic System Problems and Potentials," *A. H. Coleman, RCA Surface Communication.*

"An Investigation of the Solderless Wire Wrapped Electrical Connections," *S. Plasker, A. H. Wenser, and C. A. Selge, IBM Corp., Kingston, N. Y.*

Wednesday Afternoon

Packaging Applications for Missile and Space Vehicles

Moderator: *J. B. Duryea, GE Missile and Space Vehicle Dept.*

"High Density Electronic Packaging Via Honeycomb Structures," *C. W. Johnson, Sperry Gyroscope Co.*

"Encapsulation Techniques for Functional Blocks in Molecular Electronics," *J. M. Stiener, Westinghouse Electric Co.*

"Miniature PWM 90×10 Solid State Electronic Multiplexer for Space Vehicle Application," *A. V. Ottovino and S. G. Kritzstien, GE Missile and Space Vehicle Dept.*

Thursday Morning, June 15

Systems Engineering and Related Research and Development

Moderator: *Dr. H. Krutter, Chief Scientist NADEVCEN, Johnsville, Pa.*

"Miniature Telemetry Systems Designs," *W. C. Bennett, Teledynamics Co., Philadelphia, Pa.*

"Analysis of a Tri-Dimensional Feed Horn Support Structure," *Z. M. Slusarek, Missile and Surface Radar Div., RCA.*

"The Atlas Missile Auto Pilot—A Case History of the PGPEP Principles in Action," *J. H. Hauser, Convair Div., General Dynamics Corp.*

"Design for Future Production of the AHSR Antenna," *J. Blass and E. Auelt, W. L. Moran Co.*

Thursday Afternoon

Better Products Through Advances in Design Technology

Moderator: *Dr. A. Goldsmith, Senior Technical Consultant, RCA.*

"Practical Thermal Design Consideration for Micromodule Equipment," *Dr. P. Taylor and G. Rezek, RCA.*

"The Product Engineer and the Magnetic Thin Film," *P. Kuttner, Burroughs Corp.*

"Value Engineering and Product Design," *C. Fallon, RCA.*

"Epoxy Resin Molds for Encapsulation of Electronic Circuitry into Modules," *J. Tallent, General Electric.*

"Development of a Thermally Conductive Ceramics Component Board," *G. Kriss and L. Palaski, GE Missile and Space Vehicle Dept.*

Joint Automatic Control Conference

UNIVERSITY OF COLORADO, BOULDER, JUNE 28–30, 1961

The 1961 Joint Automatic Control Conference, June 28–30, Boulder, Colo., is the answer of five professional societies to the question of eliminating redundant meetings and providing a comprehensive forum for automatic control theory and application.

The Program Committee, chaired by M.I.T.'s Dr. H. M. Paynter, represents AIChE, AIEE, ASME, IRE and ISA (the 1961 sponsor).

Papers of all five professional societies have been combined in sessions without regard to society origin. Following a plenary session on Wednesday morning, June 28, the technical sessions are scheduled through Friday afternoon.

General Conference Chairman, R. K. Adams of Oak Ridge National Laboratories, announced that all JACC registrants will receive a complete bound digest of all papers. The essence of each paper will be presented in 750 words and up to 8 charts and illustrations. Complete papers are being reviewed for inclusion by each society in its respective *Transactions*.

Advance programs and registration forms will be available from the Meetings Manager, Instrument Society of America, 313 Sixth Ave., Pittsburgh 22, Pa.

The tentative program for the Conference has been announced as follows:

Wednesday Afternoon, June 28

Theory of Optimization

"The Structure of Optimum Control Systems," *B. Friedland.*

"On the Existence of Optimum Controls," *L. Markus and E. B. Lee.*

"Optimal Pursuit Strategies in Discrete-state Probabilistic Systems," *J. H. Eaton and L. A. Zadeh.*

"On a Property of Optimal Controllers with Boundedness Constraints," *H. L. Groginsky.*

Hydraulic and Pneumatic Control

"Design of a Hydraulic Servo with Improved Bandpass Characteristics when Driving a Resonant Mechanical Load," *W. Seamone.*

"Pneumatic Transmission Lines," *N. B. Nichols.*

"Hydraulic Control of Acoustic-Test Siren Rotors Using Analog-Digital Techniques," *F. D. Ezekiel.*

"A System Approach to High Accuracy Fluid Control Valving," *R. Henke.*

Economic Parameters in Process Control

"Cost Models for Systems Engineering," *H. Chestnut.*

"The Future Role of Feed-Forward Control in the Chemical Industry," *S. Calvert and G. Coulman.*

"Incentives for Computer Control in the Chemical Process Industries," *T. Q. Eliot and D. R. Longmire.*

Thursday Morning, June 29

Optimal Switching I

"A Switching Criterion for Certain Time-Optimal Regulating Systems," *E. E. Rang.*

"Time Optimal Control of Nonlinear Process-Equations," *E. B. Lee.*

"A General Iterative Technique for Optimal Control Systems Subject to Saturation," *Yu-Chi Ho.*

"A Minimal Time Discrete System," *C. A. Desoer and J. Wing.*

Aerospace Vehicle Control I

"Design Considerations of Inertial Wheel Systems for Attitude Control of Satellite Vehicles," *R. E. Mortensen.*

"A Decoupling Computer for Space Vehicle Attitude Control," *R. H. Cannon, Jr.*

"Derived Rate Increment Stabilization: Its Application to the Attitude Control Problem," *H. C. Vivian and J. C. Nicklas.*

Process Dynamics

"A Finite-Stage Model for Highly Asymmetric Residence-Time Distributions," *R. J. Adler and R. B. Hovorka.*

"Concentration Dynamics in Tubular Flow Systems," *L. Fan and Y. K. Ahn.*

"Transfer Functions of Heat Exchangers," *J. P. Hsu and N. Gilbert.*

Thursday Afternoon

"Theory and Design of High Order Bang Bang Control Systems," *M. Athanassiades and O. J. M. Smith.*

"Synthesis of Quasi-Stationary Optimum Nonlinear Control Systems I and II," *P. Chandsket and C. T. Leondes.*

"Optimal Control Methods for On-Off Sampling Systems," *W. L. Nelson.*

"Minimum Time Control of Second Order Pulse-Width-Modulated Sampled-Data Systems," *E. Polak.*

Aerospace Vehicle Control II

"Model Feedback Applied to Flexible Booster Control," *G. E. Tutt and W. K. Waymeyer.*

"Can Electric Actuators Meet Missile Control Requirements?" *G. C. Newton, Jr., and R. W. Rasche.*

"Terminal Control System Applications," *R. K. Smyth and E. A. O'Hern.*

Automatic Control Applications

"Direct Cycle Nuclear Power Plant Stability Analysis," *D. Buden.*

"The Application of Dead-Time Compensation to a Chemical Reactor for Automatic Control of Production Rate," *D. E. Lupfer and M. W. Oglesby.*

"Description of a Digital Speed Regulating System," *J. Dobbie and E. C. Fox.*

"Measuring and Classifying Haze in Plate Glass with the Automatic Ford Hazemeter," *B. W. Preston.*

Friday Morning, June 30

Adaptive Control Systems

"A Parameter Perturbation Adaptive Control System," *R. J. McGrath, V. Rajaraman, and V. C. Rideout.*

"Transfer Function Tracking and Adaptive Control Systems," *N. N. Pure and C. N. Weygandt.*

"Adaptive Servo Tracking," *A. I. Talkin.*

"An Adaptive Three-Mode Controller for the Process Industries," *W. B. Field.*

Nonlinear Control Systems

"A Modified Lyapunov Method for Nonlinear Stability Analysis," *D. R. Ingwerson.*

"Some Recent Advances in Analysis and Synthesis of Nonlinear Systems," *A. A. Wolf.*

"Non-Linear Controllers," *J. C. Webb and H. T. Bates.*

Statistical Design Considerations

"Signal Stabilization of a Control System with Random Inputs," *R. Oldenburger and R. Sridhar.*

"The Use of a Mean Weighted Square Error Criterion for Optimum Filtering of Nonstationary Random Processes," *K. Sahara and G. J. Murphy.*

"Stability of a Nonlinear Feedback System in the Presence of Gaussian Noise," *R. Oldenburger and R. Sridhar.*

"A General Performance Index for Analytical Design of Control Systems," *Z. V. Rekasius.*

Friday Afternoon, June 30

Special Topics in Control

"On the Periodic Modes of Oscillation in Pulse-Width-Modulated Feedback Systems," *E. I. Jury and T. Nishimura.*

"Design of Non-Interacting Control Systems Using Bode Diagrams," *K. Chen, R. A. Mathias, and D. M. Sauter.*

"A Graphical Method for Finding the Frequency Response of Nonlinear Closed-Loop Systems," *A. S. McAllister.*

"A Simple Optimizing Control for Single-Input Extremum Systems," *J. S. Frait and D. P. Eckman.*

"Sensitivity Considerations for Time-Varying Sampled-Data Feedback Systems," *J. B. Cruz.*

Process and Control Nonlinearities

"Fast Analog Computer Techniques for Design of Controllers for Nonlinear Systems," *G. J. Fiedler and J. J. Landy.*

"On Stabilization of Feedback Systems Affected by Hysteresis Nonlinearities," *A. K. Mahalanabis.*

"Stability of Servomechanisms with Friction and Stiction in the Output Element," *F. B. Tuteur.*

"The Motion of Systems Under Square Law Damping," *W. H. Baier.*

Time Series and Process Identification Process

"System Impulse Response Identification Based Upon Short Normal Operating Records," *R. Kerr and W. H. Surber, Jr.*

"Some Techniques of Linear System Identification Using Correlating Filters," *W. W. Lichtenberger.*

"Digital Computer Analysis of Closed Loop Systems Using the Number Series Approach," *R. K. Adams.*

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Broadcast & Television Receivers (G-8)—R. R. Thalner, Sylvania Home Electronics, 700 Ellicott St., Batavia, N. Y.; C. W. Sall, RCA, Princeton, N. J.

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Circuit Theory (G-4)—S. Darlington, Bell Telephone Labs., Murray Hill, N. J.; M. E. Van Valkenburg, Dept. of E.E., Univ. of Illinois, Urbana, Ill.

Communications Systems (G-19)—C. L. Engleman, Engleman & Co., Inc., 2480 16th St., N.W., Washington 9, D. C.; W. B. Jones, Jr., School of E. E., Georgia Inst. Tech., Atlanta 13, Ga.

Component Parts (G-21)—F. E. Wenger, Headquarters ARDC, Andrews AFB, Washington 25, D. C.; G. Shapiro, Engineering Electronics Sec., Div. 1.6, NBS, Connecticut Ave. and Van Ness St., Washington 25, D. C.

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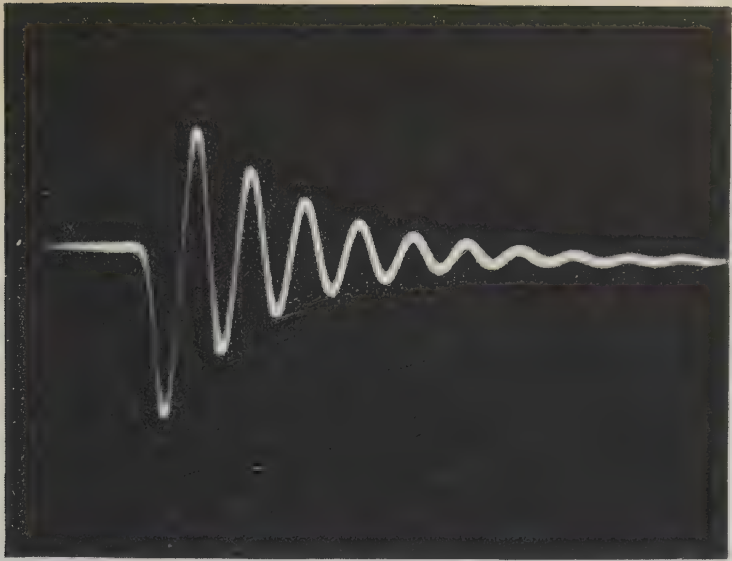

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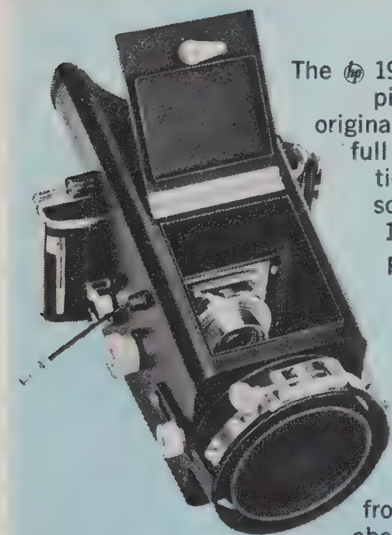

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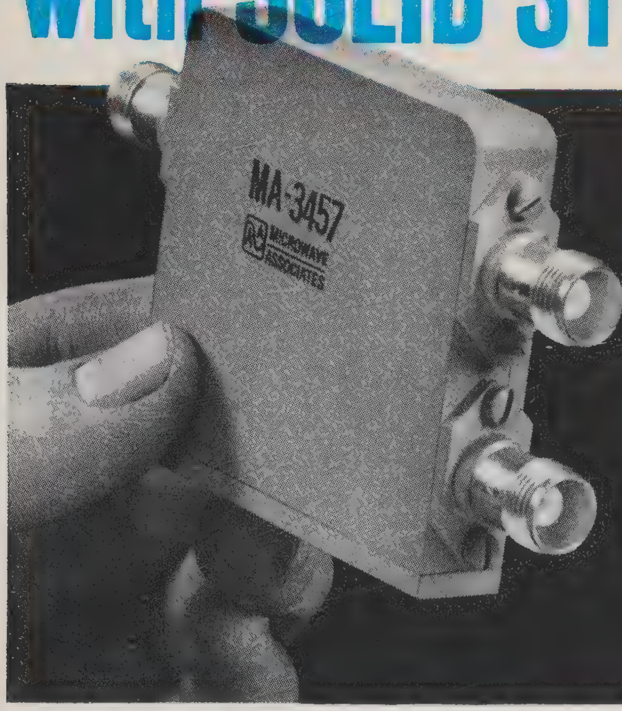

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- ★ Power handling capability { 4 watts CW (S.P.S.T. unit) / 150 watts peak at 0.001 duty cycle }
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LOW POWER LEVEL COAXIAL SWITCHES

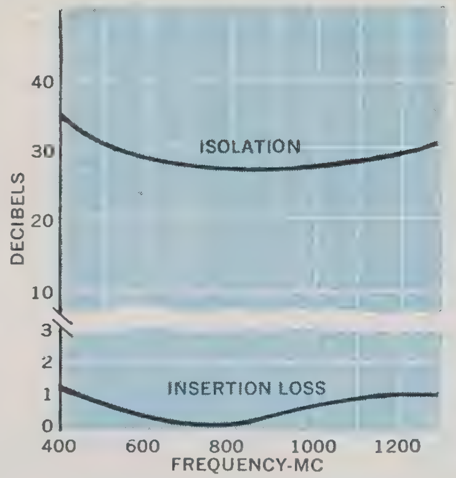
Frequency (Mc)	Insertion Loss (Max)	Isolation (Min)	Switching Power
210-240	0.2 db	20 db	10 mw
260-340	0.2 db	18 db	10 mw
400-500	0.3 db	20 db	10 mw
570-630	0.3 db	20 db	10 mw
900-1000	0.3 db	20 db	10 mw
1250-1350	0.5 db	20 db	10 mw
MEDIUM POWER LEVEL COAXIAL SWITCHES			
200-1000	1.5 db	22 db	70 mw
1000-2000	1.5 db	20 db	70 mw
2000-4000	2.0 db	16 db	70 mw

LOW POWER LEVEL VOLTAGE VARIABLE ATTENUATORS

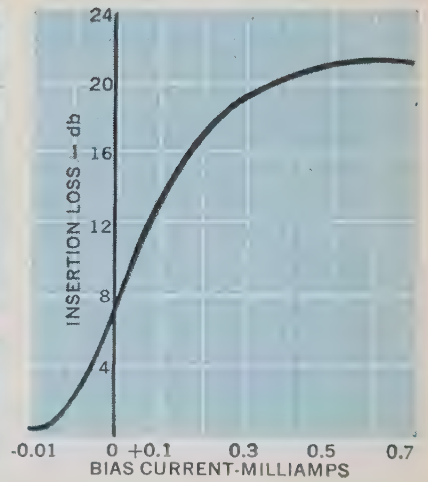
Frequency (Mc)	Attenuation Range
260-340	0.2 db-18 db
400-450	0.3 db-20 db
570-630	0.3 db-20 db
1250-1350	0.5 db-20 db

Narrow-band higher frequency units are available with lower loss and increased isolation.

TYPICAL PERFORMANCE BROADBAND MICROWAVE SWITCH SPST



TYPICAL PERFORMANCE VOLTAGE VARIABLE ATTENUATOR (425 Mc ± 25)



Units for handling higher powers are now in development. Microwave Associates has capabilities for meeting your requirements for single-pole multiple-throw and waveguide switching devices. Our switches invite comparison. We invite your inquiries. A quotation/data sheet will be sent on request.

MICROWAVE ASSOCIATES, INC.
ELECTRON TUBE AND DEVICE DIVISION, Burlington, Mass.
Western Union FAX • TWX: Burlington, Mass., 942 • BRowning 2-3000

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Rome-Utica (1)—M. P. Forte, 904 DeWitt Lane, Rome, N. Y.; J. M. Eannarino, 610 Highland Ave., Rome, N. Y.

Sacramento (7)—F. C. Jacob, Dept. of Agricultural Engrg., Univ. of California, Davis, Calif.; J. R. Kowalczyk, 3730 Las Pasas Way, Sacramento 25, Calif.

St. Louis (6)—R. D. Hill, Jr., 363 Gray Ave., Webster Groves 19, Mo.; G. E. Barnard, 639 N. 69 St., East St. Louis, Ill.

Salt Lake City (7)—C. L. Alley, Elec. Engrg. Dept., Univ. of Utah, Salt Lake City, Utah; J. E. Dalley, 3920 S. 1380 E., Salt Lake City 17, Utah.

San Antonio-Austin (6)—G. E. White, Box 9006, Allandale Station, Austin 17, Tex.; F. X. Bostick, 5002 Beverly Hills Dr., Austin 3, Tex.

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San Francisco (7)—D. A. Dunn, Stanford Electronics Labs., Stanford Univ., Stanford, Calif.; P. D. Lacy, 324 Lakeview Way, Redwood City, Calif.

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Seattle (7)—D. K. Reynolds, Dept. of Elec. Engrg., Univ. of Washington, Seattle 5, Wash.; R. M. Lundberg, 9115 N.E. 21 St., Bellevue, Wash.

Shreveport (6)—D. L. Johnson, Louisiana Polytech. Inst., Elec. Engrg. Dept., Ruston, La.; R. M. Steere, 406 West Florida Ave., Ruston, La.

South Bend-Mishawaka (5)—C. H. Hoffman, 52663 Lynnewood Ave., South Bend 15, Ind.; D. L. Cunningham, 16269 Shamrock Dr., Mishawaka, Ind.

South Carolina (3)—P. A. McMasters, 5809 Moore St., North Charleston, S. C.; H. L. Hunter, 49 Fort Dr., Rt. 6, Box 423, North Charleston, S. C.

Southern Alberta (8)—A. P. Davis, 47 Wildwood Dr., Calgary, Alta., Canada; G. L. Sadler, 2520-32 Ave., S.W., Calgary, Alta., Canada.

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Tokyo—Isaac Koga, 254 8-Chome, Kami-Meguro, Tokyo, Japan; Fumio Minozuma, 16 Ohara-Machi, Meguro-Ku, Tokyo, Japan.

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Tulsa (6)—H. B. Ferguson, 4530 E. 32 Place, Tulsa, Okla.; J. M. Bushnell, KVOO-TV, Box 1349, Tulsa, Okla.

Twin Cities (5)—J. Kahnke, 1541 Edgewater Ave., St. Paul 13, Minn.; C. G. Compton, 1011 Fairmount Ave., St. Paul 5, Minn.

Vancouver (8)—W. H. Thompson, 2958 West 28 Ave., Vancouver 8, B. C., Canada; D. H. J. Kay, 4539 Imperial St., Burnaby, B. C., Canada.

Virginia (3)—R. W. Morton, Box 96, Denbigh, Va.; J. B. Spratley, Ellerson, Va.

Washington (3)—D. C. Ports, Jansky & Bailey, 1339 Wisconsin Ave., N.W., Washington 7, D. C.; C. R. Busch, 2000 N. Vermont St., Arlington 7, Va.

Western Massachusetts (1)—J. J. Allen, 29 Sunnyside Dr., Dalton, Mass.; J. E. Mulford, Sprague Elec. Co., Marshall St., North Adams, Mass.

Western Michigan (4)—F. E. Castenholz, Police Headquarters, Jefferson & Walton Sta., Muskegon, Mich.; J. F. Giardina, 1528 Ball, N.E., R. 4, Grand Rapids 5, Mich.

Wichita (6)—M. E. Dunlap, 548 S. Lorraine Ave., Wichita 16, Kans.; N. J. Damaskos, 7803 East Indianapolis, Wichita 7, Kans.

Williamsport (5)—D. M. Jewart, 1400 Faxon Pkwy., Williamsport, Pa.; G. W. Deming, 1891 East 3rd, Williamsport, Pa.

Winnipeg (8)—P. F. Windrick, 669 Oxford St., Winnipeg 9, Man., Canada; R. I. Punshon, Canadian Broadcasting Corp., 540 Portage Ave., Winnipeg, Man., Canada.

Subsections

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Catskill (2)—E. L. Johnson, 10 Kiersted Ave., Kingston, N. Y.; C. R. Eickhorn, Jr., 10 Park Circle, Mt. Marion, N. Y.

East Bay (7)—A. J. Stripeika, 2759 Miranda Ave., Alamo, Calif.; J. T. Lavrischeff, 7029 Cutting Blvd., El Cerrito, Calif.

Eastern North Carolina (3)—W. J. Speed, 2718 E. Rothgeb Dr., Raleigh, N. C.; W. H. Horne III, Rt. 1, Raleigh, N. C.

Fairfield County (1)—T. J. Calvert, 11 Bedford Ave., Norwalk, Conn.; H. F. Wischnia, 50 De Leo Dr., Stamford, Conn.

Lancaster (3)—W. N. Parker, 1493 Hollywood Dr., Lancaster, Pa.; J. Evans, 2109 Lyndell Dr., Lancaster, Pa.

Las Cruces-White Sands Proving Ground (6)—H. Coleman, Box 1238, Las Cruces, N. M.; Secretary to be advised.

Lehigh Valley (3)—H. A. Tooker, 2524 Fairview St., Allentown, Pa.; M. C. Waltz, Bell Telephone Labs., 555 Union Blvd., Allentown, Pa.

Memphis (3)—C. Ray, Dept. of Neurosurgery, Baptist Memorial Hospital, Memphis 3, Tenn.; Brother I. J. Haas, Christian Brothers College, Memphis 4, Tenn.

Merrimac Valley (1)—C. E. White, 16 Dale St., West Peabody, Mass.; D. Christiansen, 12 Hay St., Newbury, Mass.

Mid-Hudson (2)—W. D. Reiner, IBM Corp., Dept. 553, Bldg. 703, Poughkeepsie, N. Y.; B. Augusta, 53 Colburn Drive, Poughkeepsie, N. Y.

Monmouth (2)—J. A. Young, Jr., 34 Kemp Ave., Fairhaven, N. J.; O. E. DeLange, Bell Telephone Labs., Holmdel, N. Y.

Nashville (3)—G. P. McAllister, 2923 Twin Lawn Dr., Nashville 14, Tenn.; W. B. Kincaid, Jr., 210 Graeme Dr., Nashville 14, Tenn.

New Hampshire (1)—R. Baer, 134 Mayflower Dr., Manchester, N. H.; F. J. Safford, 71 Concord St., Nashua, N. H.

Northern Vermont (1)—F. J. M. Sichel, 35 Henderson Terrace, Burlington, Vt.; W. C. Chase, WDEV, 9 Stowe St., Waterbury, Vt.

Orange Belt (7)—J. R. Mickelson, Convair, Zone 6-87, P. O. Box 1011, Pomona, Calif.; G. E. Kinzer, 2856 Ronald St., Riverside, Calif.

Palm Beach (3)—C. E. Cronin, 544 Ebb Tide Dr., North Palm Beach, Fla.; A. S. Baran, 301 Bayside Rd., Lake Worth, Fla.

Panama City (3)—S. B. Marley, 1912 Calhoun Ave., Panama City, Fla., R. C. Lowry, 2342 Pretty Bayou Dr., Panama City, Fla.

Pasadena (7)—J. W. Thatcher, 936 Winston Ave., San Marino, Calif.; R. L. Heacock, 2532 Hanning Ave., Altadena, Calif.

Pikes Peak (6)—A. O. Behnke, 204 Westcott Ave., Colorado Springs, Colo.; K. W. Linke, 07 North Garo Ave., Colorado Springs, Colo.

Reading (3)—W. I. Huyett, 1020 Wyomissing Blvd., Wyomissing, Pa.; R. H. Lundberg, 3312 Harrison Ave., Reading, Pa.

Richland (7)—P. R. Kelly, 220 Delafield, Richland, Wash.; G. L. Erickson, 213 Armistead, Richland, Wash.

San Fernando Valley (7)—J. D. Wills, 6606 Lindley, Reseda, Calif.; C. Z. Becker, RCA, 8500 Balboa Ave., Van Nuys, Calif.

Santa Ana (7)—D. R. Proctor, 1601 East Chestnut Ave., Santa Ana, Calif.; J. C. Hathaway, Collins Radio Co., 3324 West Delhi Road, Santa Ana, Calif.

Santa Barbara (7)—S. D. Crane, 590 Barker Pass Rd., Santa Barbara, Calif.; R. S. Hutcheon, 714 Chiquita Rd., Santa Barbara, Calif.

Southwestern Ontario (8)—W. A. Ruse, Bell Telephone Co., 1149 Goyeau St., Windsor, Ont., Canada; G. L. Virtue, 619 Lounsborough Rd., Sandwich-South, Windsor, Ont., Canada.

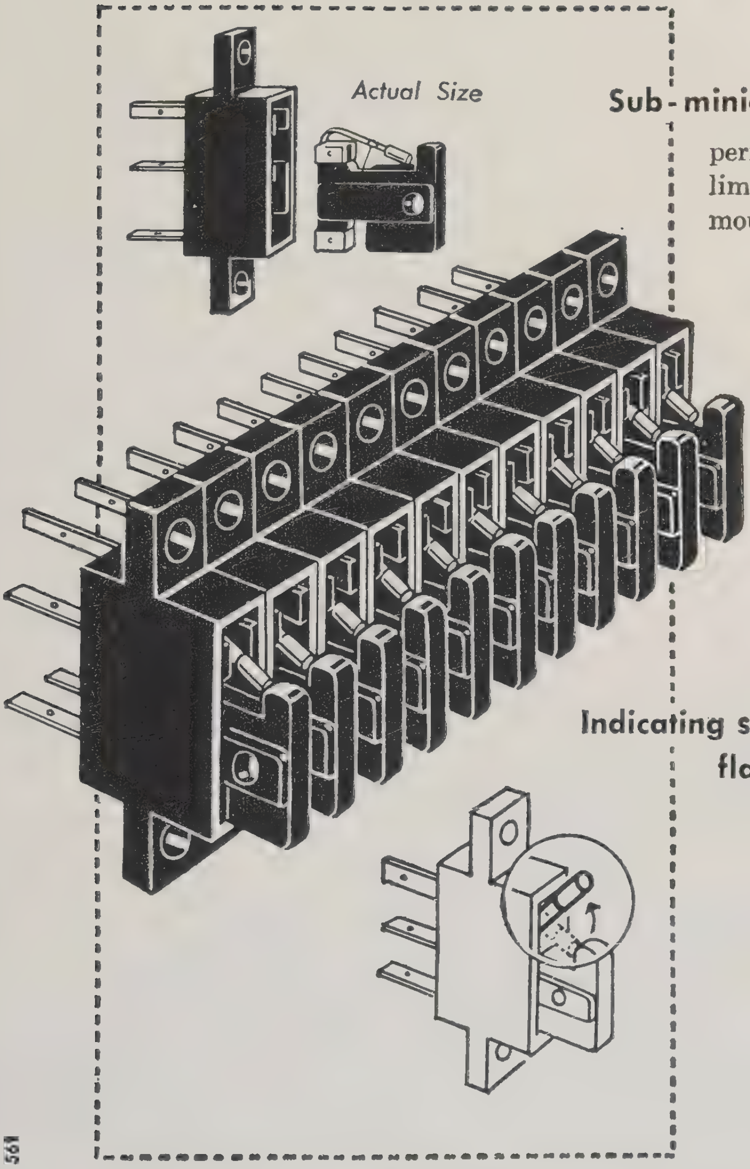
Westchester County (2)—S. Sherr, 35 Byway, Hartsdale, N. Y.; S. K. Benjamin, 59 Cooper Dr., New Rochelle, N. Y.

Western North Carolina (3)—J. I. Barron, Southern Bell T&T Co., Box 240, Charlotte, N. C.; T. C. Livingston, 926 N. Sharon Amity Rd., Charlotte, N. C.

NEW! BUSS . .

Signal Indicating · Alarm Activating

GMT Fuse & HLT Fuseholder



561

Sub-miniature design

permits multiple mounting of fuses in limited space. Fuseholders can be mounted on ¼ inch horizontal centers.

Fuse and holder combination readily adaptable for use in equipment operating at 300 volts or less, such as: communication equipment, business machines, computors, control equipment or other multiple circuit apparatus where space is at a premium.

Indicating spring flashes color-coded flag when fuse opens

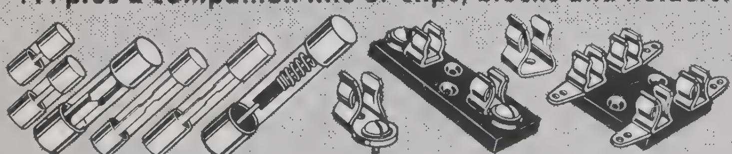
to give quick, positive identification of faulty circuit.

Indicator spring also makes contact with an alarm circuit so, it can be used to flash a light—or sound audible signal on fuse panel or at a remote location.

Ask for bulletin GMCS on BUSS GMT fuses and HLT holders.

In the BUSS line,

you'll find the type and size fuse to fit your every need . . . plus a companion line of clips, blocks and holders.



FUSETRON
TRUSTWORTHY NAMES IN ELECTRICAL PROTECTION
BUSS
McGRAW EDISON

BUSSMANN MFG. DIVISION, *McGraw-Edison Co.,* UNIVERSITY AT JEFFERSON, ST. LOUIS 7, MO.

CONVECTION COOLED
No Blowers or Filters
Maintenance Free

Highly efficient, radiator type heat sinks eliminate internal blowers, maintenance problems, risk of failure, moving parts, noise and magnetic fields. Units are rated for continuous duty at 50°C ambient.

EASY SERVICE ACCESS

Dual-deck, swing-out back construction provides simple and fast service access without the need to remove unit from rack. All major component terminals are accessible from rear.

NO VOLTAGE SPIKES OR OVERSHOOT

Lambda's design prevents output voltage overshoot on "turn on, turn off," or power failure.

MIL QUALITY

Hermetically-sealed magnetic shielded transformer designed to MIL-T-27A quality and performance. Special, high-purity foil, hermetically-sealed long life electrolytic capacitors.

SHORT CIRCUIT PROOF

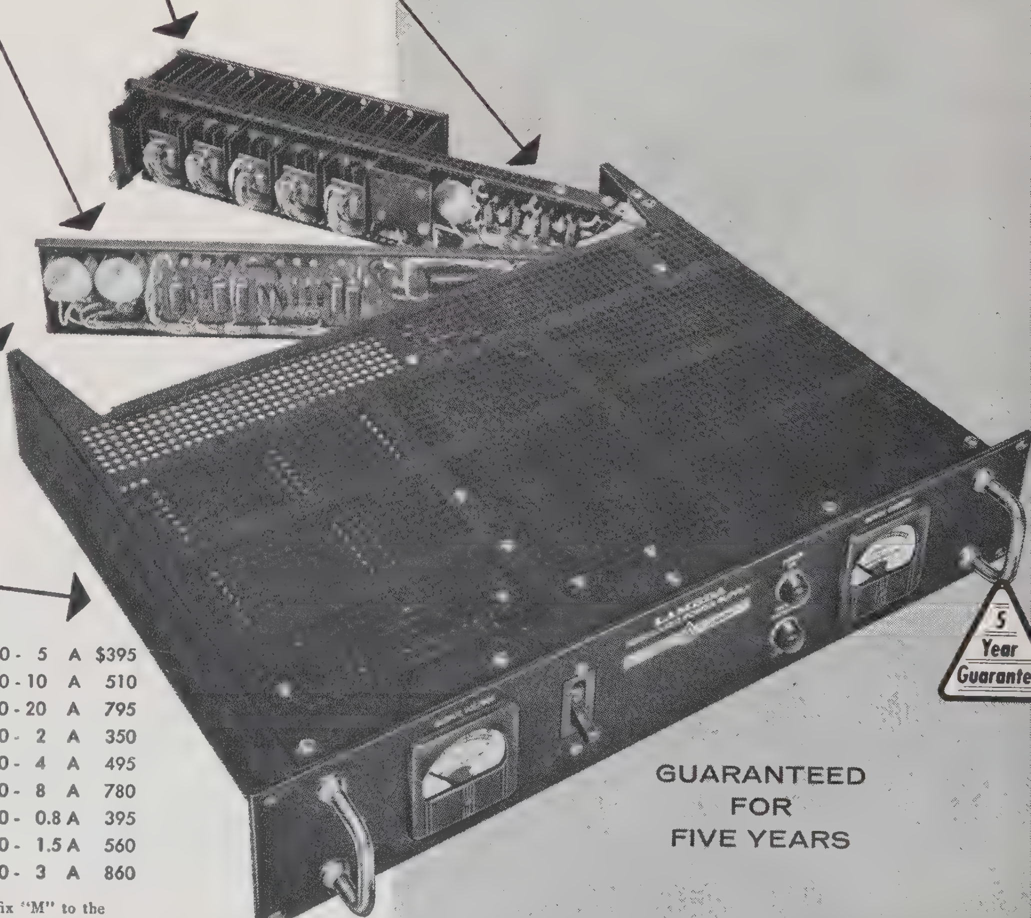
All models are completely protected with magnetic circuit breakers, fuses, and thermal overload.

REMOTE SENSING

Minimizes effect of power output leads on DC regulation, output impedance and transient response.

New
LAMBDA Transistorized REGULATED POWER SUPPLIES

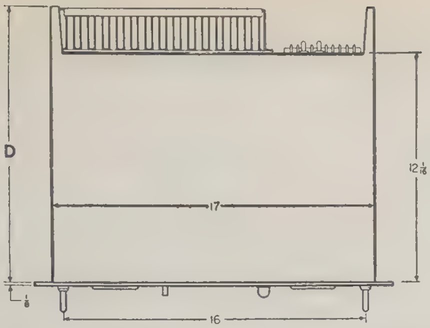
0 - 34 VDC 5, 10 and 20 Am
20 - 105 VDC 2, 4 and 8 Am
75 - 330 VDC 0.8, 1.5 and 3 Am



GUARANTEED
FOR
FIVE YEARS

LA 50 - 03A	0 - 34 VDC	0 - 5 A	$395
LA100 - 03A	0 - 34 VDC	0 - 10 A	510
LA200 - 03A	0 - 34 VDC	0 - 20 A	795
LA 20 - 05A	20 - 105 VDC	0 - 2 A	350
LA 40 - 05A	20 - 105 VDC	0 - 4 A	495
LA 80 - 05A	20 - 105 VDC	0 - 8 A	780
LA 8 - 08A	75 - 330 VDC	0 - 0.8 A	395
LA 15 - 08A	75 - 330 VDC	0 - 1.5 A	560
LA 30 - 08A	75 - 330 VDC	0 - 3 A	860

For metered models add the suffix "M" to the model number and add $30.00 to the price.



Top View
of All Models

Front View

MODEL †			
	LA 50-03A LA 20-05A LA 8-08A	LA 100-03A LA 40-05A LA 15-08A	LA 200-03A LA 80-05A LA 30-08A
A	3½″	7″	10½″
B	3″	*4″	*7½″
C	18 5/16″	18¼″	18¼″
D	14⅜″	14⅜″	16½″

* These models notched per RETMA Standards
† Includes metered models with suffix "M"

COMPLETE SPECIFICATIONS OF LAMBDA LA SERIES

DC OUTPUT (Regulated for line and load)

Model	Voltage Range (1)	Current Range	Minimum Voltage (1)	Voltage Steps (1)	Price(2)
LA 50-03A	0- 34 VDC	0- 5 AMP	0	2, 4, 8, 16, and 0- 4 volt vernier	$ 395
LA100-03A	0- 34 VDC	0-10 AMP	0	2, 4, 8, 16, and 0- 4 volt vernier	510
LA200-03A	0- 34 VDC	0-20 AMP	0	2, 4, 8, 16, and 0- 4 volt vernier	795
LA 20-05A	20-105 VDC	0- 2 AMP	20	5, 10, 20, 40, and 0-10 volt vernier	350
LA 40-05A	20-105 VDC	0- 4 AMP	20	5, 10, 20, 40, and 0-10 volt vernier	495
LA 80-05A	20-105 VDC	0- 8 AMP	20	5, 10, 20, 40, and 0-10 volt vernier	780
LA 8-08A	75-330 VDC	0- 0.8 AMP	75	15, 30, 60, 120, and 0-30 volt vernier	395
LA 15-08A	75-330 VDC	0- 1.5 AMP	75	15, 30, 60, 120, and 0-30 volt vernier	560
LA 30-08A	75-330 VDC	0- 3 AMP	75	15, 30, 60, 120, and 0-30 volt vernier	860

(1) The DC output voltage for each model is completely covered by four selector switches plus vernier control. The DC output voltage is the summation of the minimum voltage plus the voltage steps and the continuously variable DC vernier.

(2) Prices are for unmetered models. For metered models add the suffix "M" and add $30.00 to the price.

Regulation (line) ...... Less than 0.05 per cent or 8 millivolts (whichever is greater). For input variations from 100-130 VAC.

Regulation (load) ..... Less than 0.10 per cent or 15 millivolts (whichever is greater). For load variations from 0 to full load.

Transient Response..... Output voltage is constant within regulation specifications for step function:
(line) .... line voltage change from 100-130 VAC or 130-100 VAC.
(load) .... load change from 0 to full load or full load to 0 within 50 microseconds after application.

Internal Impedance .... LA 50-03A less than .008 ohms
LA100-03A less than .004 ohms
LA200-03A less than .002 ohms
LA 20-05A less than .06 ohms
LA 40-05A less than .03 ohms
LA 80-05A less than .015 ohms
LA 8-08A less than .5 ohms
LA 15-08A less than .25 ohms
LA 30-08A less than .15 ohms

Ripple and Noise ...... Less than 1 millivolt rms with either terminal grounded.

Polarity .... Either positive or negative terminal may be grounded.

Temperature Coefficient .... Less than 0.025 %/°C

AC INPUT.............. 100-130 VAC, 60 ± 0.3 cycle[3]
LA 50-03A....360 watts[4]
LA100-03A....680 watts[4]
LA200-03A...1225 watts[4]
LA 20-05A....390 watts[4]
LA 40-05A....710 watts[4]
LA 80-05A...1350 watts[4]
LA 8-08A....415 watts[4]
LA 15-08A....760 watts[4]
LA 30-08A...1450 watts[4]

[3]*This frequency band amply covers standard commercial power lines in the United States and Canada.*
[4]*With output loaded to full rating and input at 130 VAC.*

AMBIENT TEMPERATURE AND DUTY CYCLE Continuous duty at full load up to 50°C (122°F) ambient.

OVERLOAD PROTECTION:
Electrical .... Magnetic circuit breaker front panel mounted. Special transistor circuitry provides independent protection against transistor complement overload. Fuses provide internal failure protection. Unit cannot be injured by short circuit or overload.
Thermal .... Thermostat, manual reset, rear of chassis. Thermal overload indicator light front panel.

METERS .... Voltmeter and ammeter on metered models.

CONTROLS:
DC Output Controls ... Voltage selector switches and adjustable vernier-control rear of chassis.
Power .... Magnetic circuit breaker, front panel.
Remote DC Vernier .... Provision for remote operation of DC vernier.
Remote Sensing .... Provision is made for remote sensing to minimize effect of power output leads on DC regulation, output impedance and transient response.

PHYSICAL DATA:
Mounting .... Standard 19″ Rack Mounting

Size
LA 50-03A, LA20-05A, LA 8-08A 3½″ H x 19″ W x 14⅜″D
LA100-03A, LA40-05A, LA15-08A 7″ H x 19″ W x 14⅜″D
LA200-03A, LA80-05A, LA30-08A 10½″ H x 19″ W x 16½″D

Weight
LA 50-03A, LA20-05A, LA 8-08A 55 lb Net 85 lb Ship. Wt.
LA100-03A, LA40-05A, LA15-08A 100 lb Net 130 lb Ship. Wt.
LA200-03A, LA80-05A, LA30-08A 140 lb Net 170 lb Ship. Wt.

Panel Finish .... Black ripple enamel (standard). Special finishes available to customers' specifications at moderate surcharge. Quotation upon request.

IRE People

The appointment of **Dr. Leonard G. Abraham, Jr.** (S'49–A'54) as an engineering specialist at the Applied Research Laboratory of Sylvania Electric Products Inc., has been recently announced.

L. G. Abraham, Jr.

The Applied Research Laboratory is the central research facility for Sylvania Electronic Systems, and has the responsibility for the investigation and the development of new electronic systems and techniques to meet the requirements of government and industry.

Prior to joining Sylvania, Dr. Abraham was a research associate with General Electric Research Laboratory in Schenectady, N. Y., where he was engaged in research on radio wave propagation, character recognition, switching and information theory. While there, he also participated in a three-year study in tropospheric scatter propagation under the sponsorship of the U. S. Air Force Cambridge Research and Development Center

A naval veteran of World War II, he received the Bachelor's, Master's and Doctor's degree in electrical engineering from Cornell University, Ithaca, N. Y. He is a member of Eta Kappa Nu, Tau Beta Pi, Sigma Xi, Phi Kappa Phi, honorary scientific fraternities, and a former Fellow in the National Science Foundation.

❖

Telechrome Manufacturing Corp., Amityville, New York, has opened a Washington, D. C. office, in charge of **Robert Adams** (M'57–SM'59), newly-appointed manager of military products for the company.

R. Adams

He was formerly manager of eastern operations for Packard-Bell Electronics Corporation, and will conduct military liaison for Telechrome's Electronics and Hammarlund-Automation Divisions, the Hammarlund Manufacturing Company, and Universal Transistor Products Corporation.

Mr. Adams is a native of Philadelphia, Pa., a graduate of Drexel Institute of Technology in that city, and a member of the Quarter Century Wireless Association.

❖

Hal Borgen (A'60–M'61), formerly sales manager of Lumen, Inc., has joined Instruments for Industry, Inc., of Hicksville, N. Y., as manager of product development.

In addition to component development and sales for IFI, he will be responsible for sales of the company's new Rattray division, which designs and manufactures high precision potentiometers.

He entered industry 10 years ago after serving on the economics faculties of New York University, New York, N. Y., and the University of Nebraska, Lincoln. During World War II he was with the Army Signal Corps and served with a radar group in the Pacific. After the surrender of Japan, he set up troop training courses in what became known informally as the "College of the Marianas."

He attended Crane College, Chicago-Kent College of Law and the University of Chicago before entering the Service. He completed his education as a graduate student at New York University and the University of Nebraska after the war.

He is a member of the American Institute of Electrical Engineers, the Instrument Society of America and a charter member of the Electronics Sales Managers Association.

❖

Dynamics Corporation of America has announced establishment of a European Division, with headquarters in London, England, and appointment of **W. Allen Bridges** (M'60), electronics and communications specialist, to head it as Director of European Operations.

W. A. Bridges

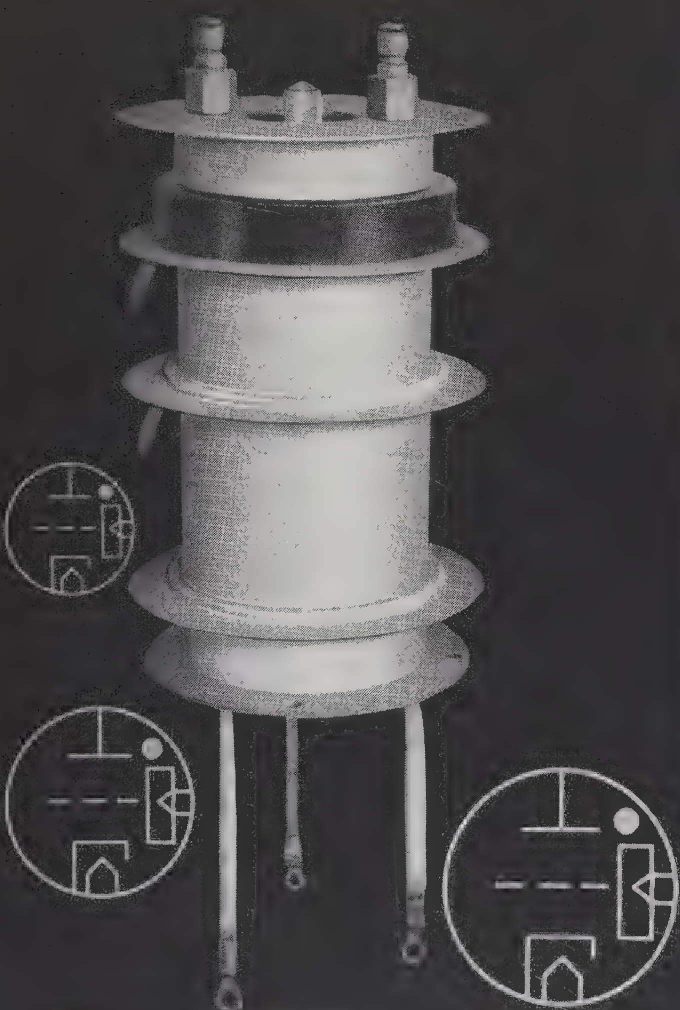
He will have over-all responsibility for the company's European activities and its dealings with private firms and governmental agencies, and he will be assigned to the Boards of Directors of present and future members of the DCA "European Group." He has been elected to the Board of Winston Electronics. His duties will also include liaison between various DCA "sister companies" in Europe, as these are formed, and representing the interests in Europe of DCA American subsidiaries and divisions.

He joined DCA in January, 1961, and previously had been European consultant to a number of American companies, including DCA's communications subsidiary, Radio Engineering Laboratories, Inc. During this period (1957–1961), he had resided in France. Prior to establishing his

(Continued on page 35A)

Progress In Hydrogen Thyratrons

55×10^9 ANODE DISSIPATION FACTOR



G-E Power Tube Department Products Also Include:

- Ignitrons
- Traveling-wave tubes
- Magnetrons
- Metal-ceramic tetrodes
- High-power duplexers
- High-power waveguide filters
- Klystrons
- Thermionic converters

Highest Registered Rating Now Available from G.E. In an Air-cooled Tube

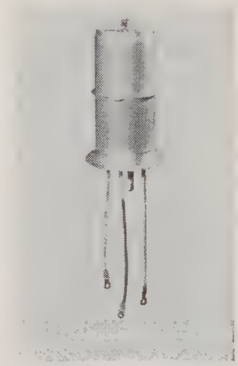
The latest addition to General Electric's expanding line of hydrogen thyratrons is now available for pulse applications such as radar modulators and linear accelerators. Developed under U. S. Army Signal Corps contract, the GL-7890 achieves an anode dissipation factor of 55×10^9 and has a peak anode voltage rating of 40 kv. The tube can now be operated water-cooled or air-cooled at full ratings.

COMING: INCREASED CURRENT AND VOLTAGE CAPACITY

Now in the late stages of development, the Z-5212 will further increase voltage and current-carrying capacity in hydrogen thyratrons. Peak anode voltage rating for this tube will be 50 kv with an average current rating of 8 amp. General Electric's Power Tube Department will welcome your requests for technical data on the Z-5212.

TEMPERATURE INDICATING DEVICE ON GL-7390A

The first high-power ceramic-metal hydrogen thyratron, General Electric's GL-7390, is now being built to MIL specifications. A modified version of this tube, the GL-7390A, is equipped with an integral anode temperature indicator for convenient readings. Both the GL-7390 and the GL-7390A have ratings of 33-kv peak anode voltage and 4-amp average current.



GL-7390A

HYDROGEN THYRATRON BULLETIN AVAILABLE

For a comprehensive analysis of the theory and application of hydrogen thyratrons, write to the Power Tube Department, General Electric Company, Schenectady, N. Y. Ask for Bulletin PT-49. To order, or obtain more information on hydrogen thyratrons, contact your nearest Power Tube sales office. Phone numbers are listed below.

265-09-9545-8481-36

POWER TUBE DEPARTMENT

GENERAL ELECTRIC

TELEPHONE TODAY—Syracuse, OL 2-5102 . . . Clifton, N. J., GR 3-6387 . . . New York City, WI 7-4065 . . . Washington, D. C., EX 3-3600 . . . Chicago, SP 7-1600 . . . Dayton, BA 3-7151 . . . Orlando, Fla., GA 4-6280 . . . Los Angeles, GR 9-7765.

(Continued from page 32A)

own consultant firm, he had worked for major American electronic companies, including Motorola, Inc., Lockheed Aircraft Corporation and Thompson Ramo-Wooldridge Corporation, in the missiles systems and communications fields.

A graduate of the College of William and Mary, Williamsburg, Va., where he received the B.S. degree in Physics, and was elected to Phi Beta Kappa, Mr. Bridges also studied at the School of Industrial Management, Massachusetts Institute of Technology, Cambridge. He served in World War II in the U. S. Army Signal Corps.

❖

Dr. S. F. Crumb (SM'55) has been appointed Professor of Electrical Engineering at Arlington State College, Arlington, Tex. He will teach and do research and consulting in the fields of networks, servomechanisms, and computers. He was formerly head of the Navigation and Guidance Systems Group at Convair, Fort Worth, Tex., a division of General Dynamics Corporation. He received the B.S. and M.S. degrees from the University of Texas, Austin, and the Ph.D. degree from the California Institute of Technology, Pasadena, in 1954. He has served as an instructor at the University of Texas and at the California Institute of Technology, and was a graduate lecturer at Southern Methodist University, Dallas, Tex.

His industrial experience includes that of test engineer at General Electric Company (1943–1944); engineer with the Magnolia Pipe Line Company (1947–1948); and research engineer with the Defense Research Laboratory (1949–1950). In 1954 he joined Convair as a Senior Aerophysics Engineer, advancing to the rank of Assistant Project Engineer.

Dr. Crumb is a member of Eta Kappa Nu, Tau Beta Pi, Sigma Xi, AIEE, and is a registered professional engineer in the state of Texas.

❖

William W. Drake, Jr. (A'52–M'55), Vice President—Administration of Aerospace Corporation, has been elected by the board of trustees to the additional position of Treasurer.

W. W. Drake, Jr.

The 39 year-old executive was named Vice President in August, 1960, after serving as a consultant during the organization of Aerospace Corporation, which serves as the Air Force's technical arm in ballistic missile and space programs.

Prior to joining Aerospace, he was As-

(Continued on page 36A)

(Continued from page 35A)

sistant to the Vice President for Europe of the Raytheon Company. Previously he was Assistant to the Director, Continental Test Operations, of the Los Alamos Scientific Laboratory, Los Alamos, N.M.

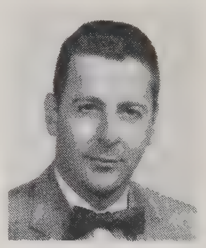
From 1947 to 1949, he was Research Associate at the Laboratory of Nuclear Science and Engineering at Massachusetts Institute of Technology, Cambridge. Earlier he taught chemistry for a year at the University of Massachusetts, Amherst.

A native of New York, N. Y., he attended high school at Freehold, N. J., and in St. Louis (Principia Upper School). He received the B.S. degree in physics from Principia College at Elsah, Ill., and later studied at the Harvard Graduate School of Arts and Sciences, Cambridge.

Mr. Drake served as an electronics officer in the U. S. Navy during World War II. He is a member of the American Physical Society.

❖

Promotion of **William H. Fahringer** (A'47) to Mass Memory Systems Manager, Bryant Computer Products Division, Ex-Cell-O Corporation, has been announced.



W. H. Fahringer

Before joining Bryant as Senior Development Engineer last year, he served with the Raytheon Company, Newton, Mass., as Resident Engineering Supervisor at the Instrumentation Laboratory of the Massachusetts Institute of Technology, Cambridge. He was responsible for development of magnetic memory elements for the new 2,500 mile range Polaris Missile.

He entered the computer field with Burroughs Corporation, Detroit, Mich., as Senior Digital Data Processing Engineer in 1953. He had formerly spent nine years as Broadcast Engineer with WJR Radio, Detroit.

Mr. Fahringer is a native of Bridgeport, Conn.

❖

Tudor R. Finch (M'46) has been named Assistant General Manager of Motorola's Semiconductor Products Division, it was recently announced.

T. R. Finch

Previously, he had been a Special Systems Development Engineer at Bell Telephone Laboratories, Murray Hill, N. J., where he was responsible for research in the field of electronic switching with particular emphasis on de-

(Continued on page 38A)

(Continued from page 36A)

velopment of solid state logic and storage circuits for a time division switching system.

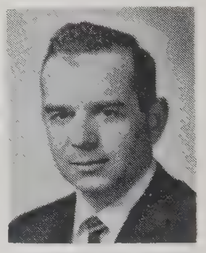
In his new position, he will have responsibility for over-all coordination of research and development of new product lines. In addition he will supervise the Division's professional recruiting and training programs and will serve as a staff assistant to the vice president and general manager of the Semiconductor Division.

He received the B.S. degree in electrical engineering in 1938 and the M.S. degree in 1939, both from the University of Colorado, Boulder. He has been a very active member of technical committees of the IRE, such as Solid State Electronics. He was named 1961 chairman for the International Solid State Circuits Conference, which was held February 14–17 in Philadelphia, Pa.

During World War II, Mr. Finch participated in the development of precision radar systems and components such as fractional microsecond pulse measuring equipment, wideband, pulse forming and delay networks. In 1952 he became head of a group engaged in exploratory transistor circuit development for digital and analog computers.

⁂

Dr. Morris Handelsman (SM'55) has been promoted to the position of Associate Director of Advanced Military Systems, Defense Electronic Products, Radio Corporation of America, it was recently announced.



M. HANDELSMAN

For the past two years, he has been a member of the Technical Staff of AMS, an organization established by RCA at Princeton, N. J., to create and develop new and advanced weapon systems concepts. Prior to his promotion, his work was in the field of Undersea Warfare at AMS.

Before joining Advanced Military Systems, he was Associate Director of Research and Development at Rome Air Development Center, Griffiss Air Force Base, Rome, N. Y. Here he was concerned with development of systems in radar, countermeasures and communications.

A native of New York, N. Y., he holds the B.S. degree in electrical engineering from the City College of New York, the Master's degree from Ohio State University, Columbus, and the Ph.D. degree in electrical engineering from Syracuse University, Syracuse, N. Y.

Following wartime service as an officer in the U. S. Army Signal Corps, he became associated with Air Force research and development at Wright Field, working in the area of radar components and techniques.

(Continued on page 40A)

IRE People

(Continued from page 38A)

In 1953 he was appointed Chief of the Radar Laboratory at the Rome Center with responsibility for research, development and early production activities for the Air Force ground radar systems. In February, 1956, he became Associate Director of Research and Development at Rome, a position which he held until joining RCA in June, 1959.

Dr. Handelsman is a member of the Acoustical Society of America and Sigma Xi. He is author of technical papers in the fields of microwaves, antennas, radar systems, satellite communications and sonar systems. He was a lecturer in electrical engineering at Syracuse University, giving graduate courses in systems engineering, antennas, and radar.

❖

Jack R. Hauser (A'54) has been named general sales manager of California Technical Industries, a division of Textron, Inc.

J. R. Hauser

In his new post, he will direct sales activities for the company's total product line.

He was formerly with Ampex Corporation for a period of eight years, most recently in marketing and sales administration with Ampex Professional Products Company, Redwood City, Calif. Prior to this, he served as a sales engineer with Minnesota Mining and Manufacturing Company for three years. Earlier he was associated with National Cash Register Company and U. S. Rubber Company.

Mr. Hauser was graduated from Stanford University, Stanford, Calif., in 1945 with an A.B. degree in mechanical engineering, and also attended the California Institute of Technology, Pasadena. He is a member of the Society of Motion Picture and Television Engineers, and other professional organizations.

❖

Herbert A. Finke (M'43) has been named to the new position of Vice President and General Manager of Bomac Laboratories, Inc.

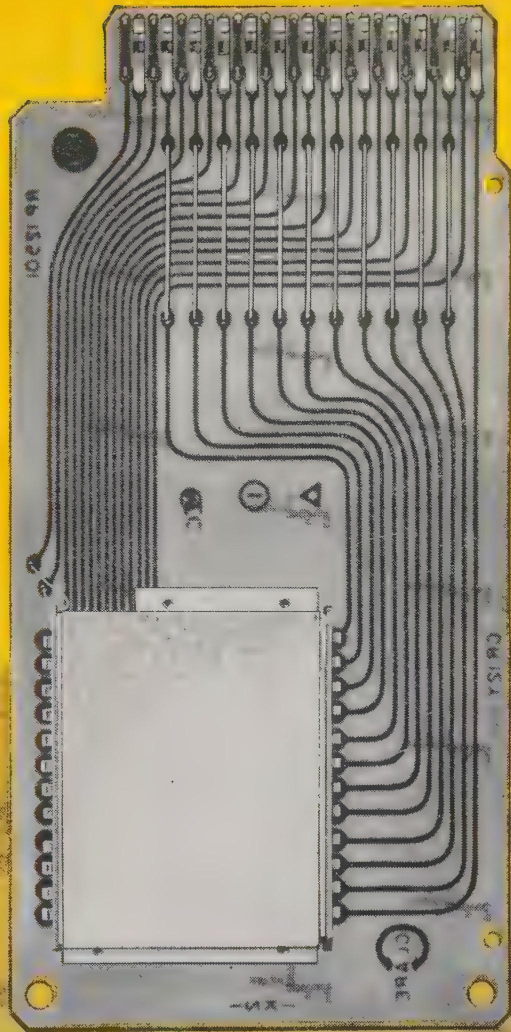
He will assume his new duties immediately, leaving his position as director of long-range planning for Varian Associates of Palo Alto, California, parent company of Bomac.

A native of New York, N. Y., he joined Varian in 1960. Prior to that he was Vice President and General Manager of Polytechnic Research and Development Corporation, a subsidiary of Harris-Intertype Corporation of Cleveland, Ohio. He has also held positions with RCA and United Aircraft.

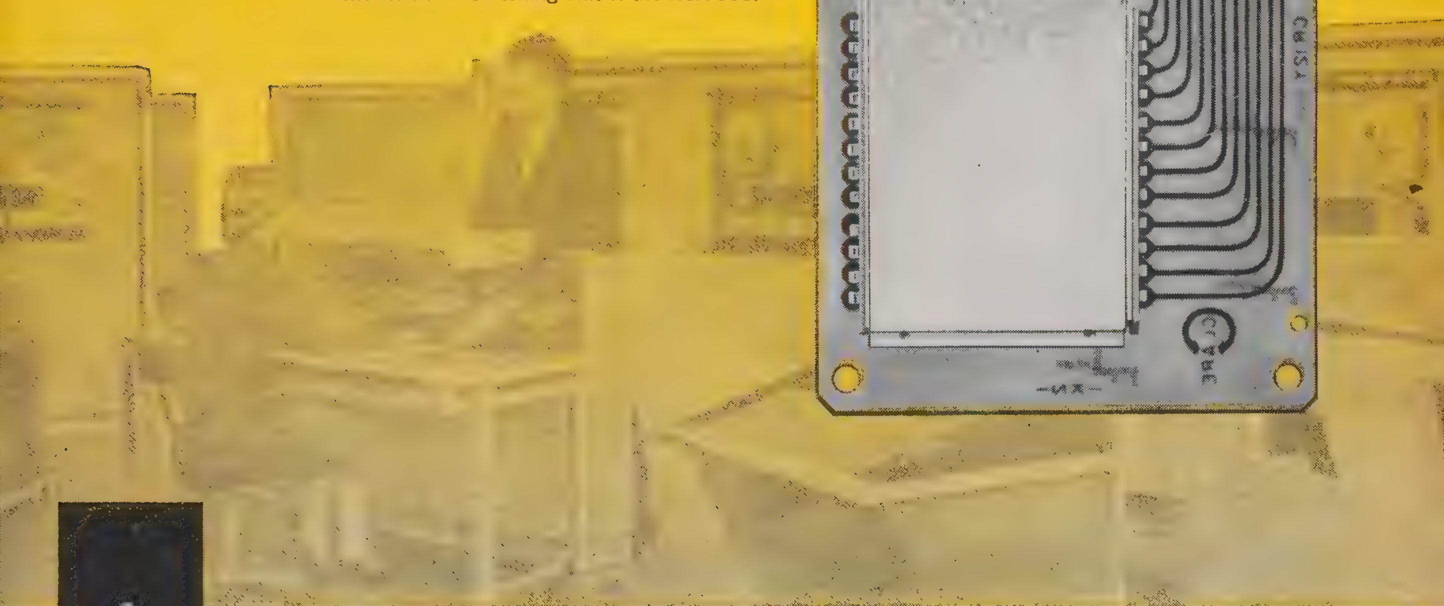
(Continued on page 43A)

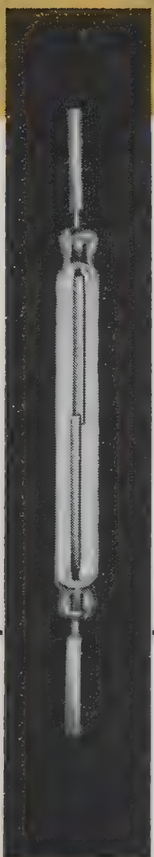
RCA uses 252 CLARE Printed Circuit Relays in the 501 electronic data processing system

RCA's 501 incorporates many advanced features which significantly increase reliability as well as economy. It takes up less space, weighs less and operates on less electrical power than previous models.



252 relays (each consisting of 12 Clareed sealed contact reed switches—3,024 switches in all) make up this "matrix relay," used in the model 547-6 switching unit of the RCA 501.





CLAREED Sealed Contact Relays provide fast, sure switching

Contributing to the efficiency, speed and compact structure of the RCA 501 are 252 CLAREED sealed contact reed relays. Mounted on printed circuit boards, these relays, their contacts hermetically sealed in contaminant-free inert gas, assure millions of perfect operations...hundreds of millions when operated at up to ½ rated load.

CLAREED relays are ideal components for transistor-drive applications such as the RCA 501. Their low inductance, and the low inductance change in the operating coil at each operation, limit the transients produced.

These relays may be mounted to meet the requirements of almost any application or environment. Consult your nearby CLARE sales engineer...or write: C. P. Clare & Co., 3101 Pratt Blvd., Chicago 45, Illinois. In Canada: C. P. Clare Canada Ltd., 840 Caledonia Road, Toronto 19, Ontario. Cable Address: CLARELAY. Ask for Bulletin CPC-10.

CLAREED switch capsule consists of a pair of magnetically operated contacts, hermetically sealed in an atmosphere of inert gas.

C. P. CLARE & CO.

Relays and Related Control Components

IRE People

(Continued from page 40A)

A graduate of Massachusetts Institute of Technology, Cambridge, he received the Master's Degree in electrical engineering from Polytechnic Institute of Brooklyn, Brooklyn, N. Y. He is the author of a number of scientific papers and holds some 10 U. S. patents for electronic devices.

Mr. Finke is a member of the American Management Association, Phi Beta Kappa, and Sigma Xi.

❖

Peter P. Grad (A'57) has been appointed Chief Chemical Engineer for Rotron Manufacturing Company, Inc., according to a recent announcement.

P. P. Grad

He holds the Bachelor of Science degree in chemistry from Polytechnic Institute of Brooklyn, Brooklyn, N. Y., and a diploma in chemical engineering from Pratt Institute, Brooklyn, N. Y. Prior to joining Rotron, he served as Chief Research Chemist and Technical Director at Aerovox Corporation, and Chief Analyst in the Chemical Division of the Borden Company.

Mr. Grad's professional memberships include the American Chemical Society, Society of Plastics Engineers, Electromechanical Society, National Academy of Sciences, and the National Research Council. He is a former member of the Board of Trustees, New Bedford Institute of Technology, and the South Eastern Massachusetts Institute of Technology.

❖

Clyde H. Hoffman (A'54–M'57–SM'59), Assistant Professor of Electrical Engineering at the University of Notre Dame, Notre Dame, Ind., addressed the Western Michigan Section of the IRE at its meeting held at Western Michigan University in Kalamazoo, Mich., on March 13, 1961. The occasion was the chartering of a new Student Branch of the Institute at Western Michigan University.

C. H. Hoffman

Professor Hoffman, Chairman of the Region V Education Committee of the Institute, spoke on the functions and activities of the Institute and of the responsibilities of the members to their Society and the Profession.

❖

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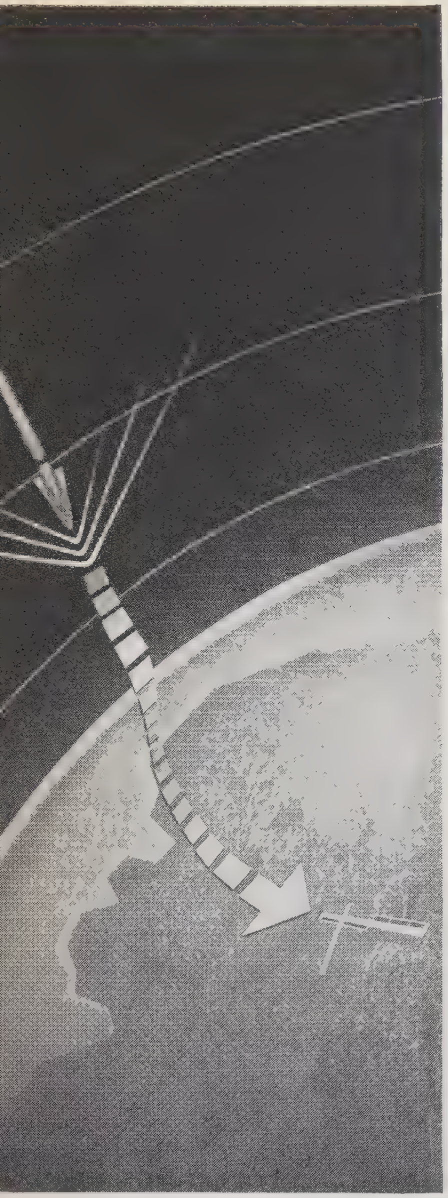

HIGH-G
INERTIAL
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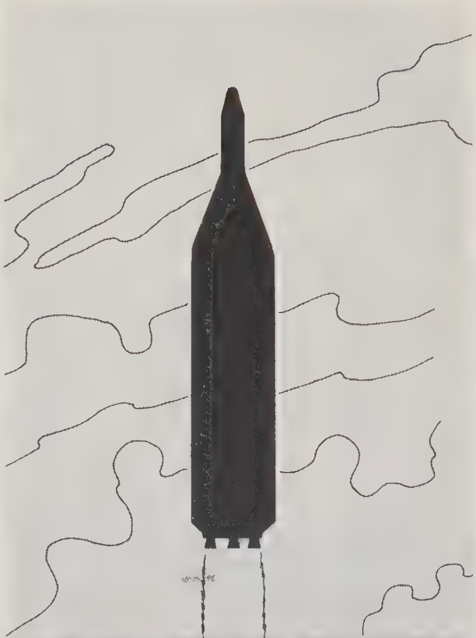
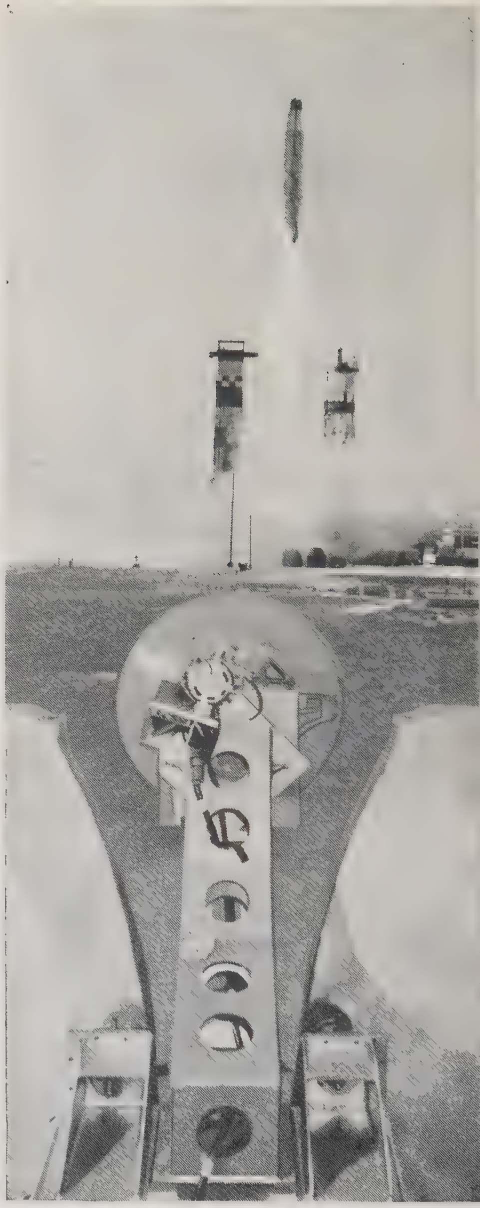
GUIDANCE



FIRE CONTROL



ANTENNAS

RESEARCH, DESIGN AND PRODUCTION OF PRECISION ORDNANCE EQUIPMENT—SINCE 1941

IRE People

(Continued from page 43A)

Dr. R. K. Hellmann (A'38–VA'39–M'50–SM'59) of the Hazeltine Corporation has been named a Fellow of the American Institute of Electrical Engineers "for contributions in the field of civilian and military electronics."

A Vice President and Director of Hazeltine Research Corporation, a Vice President of Wheeler Laboratories, Inc., and an Assistant Vice President of Hazeltine Electronics Division, Dr. Hellmann has been responsible for important research and development work at Hazeltine since 1947. He has several patents to his credit and is the author of numerous articles. He is also a past Chairman of the Long Island Chapter of the IRE.

❖

Dr. C. W. Jiles (M'55–SM'59) has recently joined the staff of Arlington State College, Arlington, Tex., as a professor of electrical engineering. He will teach and consult in the fields of circuits, servomechanisms, and network synthesis.

C. W. Jiles

He received the B.S. and B.A. degrees in electrical engineering and mathematics from the Louisiana Polytechnic Institute, Ruston, in 1949, and the M.S. and Ph.D. degrees from Oklahoma State University, Norman, in 1950 and 1955, respectively. His teaching experience includes instructor at Oklahoma State University and graduate lecturer at Southern Methodist University, Dallas, Tex. From 1955 to 1960 he was with Convair, Fort Worth, Tex., as a specialist in servomechanisms and automatic control systems. He began as Senior Aerophysics Engineer and advanced to the rank of Design Specialist.

Dr. Jiles is a member of Eta Kappa Nu, Sigma Pi Sigma, Phi Kappa Phi, AIEE, and is a registered professional engineer in Texas and Oklahoma.

❖

Raymond Kendall (A'41–M'45–SM'57) has been appointed manager of research and engineering programs at Information Technology Laboratories, a division of Itek Corporation.

R. Kendall

He is a graduate of Case Institute of Technology, Cleveland, Ohio, and has taken special courses in marketing and managerial economics at the University of Chicago, Chicago, Ill.

(Continued on page 48A)

IRE People

(Continued from page 46A)

He is a registered professional engineer in Ohio, and holds offices in the American Institute of Electrical Engineers.

He is also Vice Chairman of the Professional Group on Engineering Writing and Speech of IRE. A member of the Alumni Council at Case Institute of Technology, he belongs to the National Pilot's Association and the Bay State Flying Club. He has had several articles published and presents papers frequently.

Coming to ITL from Raytheon, Mr. Kendall has also worked for Motorola, Inc., Collins Radio Company and Kendall Radio Company.

❖

The appointment of **Ralph G. Lindstrom** (A'54–M'59–SM'59) as Manager, Engineering Services Department for the Western Development Laboratories of Philco Corporation, was recently announced.

A native of Ohio, he received the B.S. degree in electrical engineering from the Ohio Northern University, Ada.

In this new post he will be responsible for directing the activities of the Technical Standards Section, Data Control Section, Drafting and Design Section, and Components Section at the Western Development Laboratories.

A veteran in Engineering Standards and Components, he has been associated with Farnsworth Electronics Company and Associated Missile Products Corporation. Prior to joining Philco he was Supervisor of Engineering Services for Lockheed Aircraft Service, Ontario, Calif.

He is a member of the IRE Professional Group on Components and of the Standards Engineers Society.

❖

Polytechnic Institute of Brooklyn, Brooklyn, N. Y., has announced the appointment of **Dr. Nathan Marcuvitz** (S'36–A'37–VA'39–M'55–SM'57–F'58) as Vice President for Research.

He has been director of Polytechnic's Microwave Research Institute since 1957, as well as supervisor of its electrophysics group since 1952. He has also been electrophysics consultant to New York University's Institute of Mathematical Sciences for the last 12 years.

In his new position, the 48-year-old electrical engineer will direct research projects totaling more than $3,500,000 in annual expenditures. Polytechnic research is carried on in the fields of aerospace engineering, electrical engineering, chemistry, chemical engineering, physics, mechanical engineering, civil engineering, metallurgical engineering and mathematics.

Dr. Marcuvitz received the bachelor's, master's and doctorate degrees from Polytechnic. While a graduate student, he worked as a development engineer for the Radio Corporation of America (1936–40), and served as a Fellow at Polytechnic.

(Continued on page 50A)

(Continued from page 48A)

During World War II he was a research associate at the Radiation Laboratory of the Massachusetts Institute of Technology, Cambridge (1941–46).

Author of the "Waveguide Handbook," he is a member of the American Physical Society, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

❖

Louis Martin (M'44–SM'47) has joined Eitel-McCullough, Inc., San Carlos, Calif., as Manager of Marketing Operations. He will direct commercial, government and export marketing, advertising and sales promotion, and market research.

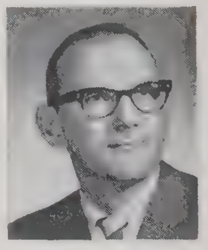
L. MARTIN

He brings 26 years of electronics industry experience to his new position at Eimac. Recently General Marketing Manager of the Westinghouse Electron Tube Division at Elmira, N. Y., he was earlier with RCA for fifteen years as Engineer, Application Engineering Manager, and Equipment Field Force Manager.

Mr. Martin received the Bachelor's Degree in electrical engineering from Cooper Union, New York, N. Y., and did graduate work at Columbia University, New York, N. Y., and the University of Pennsylvania, Philadelphia. He has served on several committees of the Electronics Industry Association.

❖

The appointment of **Robert P. McGaughey** (M'56) as manager of Computer Measurements Company's new Systems Engineering Department has been announced.



R. P. MCGAUGHEY

He comes to the company with more than 20 years experience in electronics instrumentation and communications. His most recent post was at Hoffman Electronics, where he had project responsibility for a Passive Countermeasures System designed for the Navy. The system includes a 100-megacycle counter and frequency divider.

Previously, Mr. McGaughey was engineer in charge at Radioplane Company for the design and development of a television reconnaissance system for the Signal Corps. From 1944 to 1956 he held several positions with the American Broadcasting Company.

He is a member of the Society of Motion Picture and Television Engineers.

❖

(Continued on page 52A)

IRE People

(Continued from page 50A)

Hyman Rosen (M'58) has been appointed chief electronic engineer of the new Ground Support and Systems Engineering Division of Tenney Engineering, Inc., Union, N. J.

H. Rosen

He was formerly with Flight Support, Inc. He is a graduate of the Newark College of Engineering, Newark, N. J., and is a member of the American Institute of Electrical Engineers.

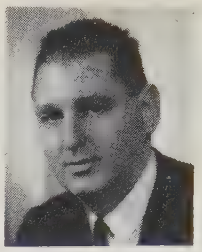
❖

At a recent meeting of the Board of Directors, **L. Dennis Shapiro** (S'52–M'57) was elected President of Aerospace Research, Inc., Cambridge, Mass.

Mr. Shapiro, who is also director of the research activities at ARI was formerly Chief, Space Research Section, Pickard & Burns, Inc., Needham, Mass. He has been involved in such programs as the International Geophysical Year, Hardtack, Argus, Firefly, and has conducted research in nuclear effects and over-the-horizon missile detection.

❖

The Nuclear Corporation of America has announced the appointment of **John S. Rydz** (M'60) as Executive Vice President. He formerly was manager of special corporate projects for the Radio Corporation of America.



J. S. Rydz

He was a member of the staff of RCA's Senior Executive Vice President. He also initiated and directed development projects in cryogenics, masers, millimeter waves, molecular frequency standards, nuclear environment, electrostatic printing, plasma physics, and flying spot scanning systems.

Before joining RCA in 1952, he held positions at the Massachusetts Institute of Technology's Color Laboratory, and at Raytheon Company in instrumentation and control areas. During World War II he worked on electronics projects while serving with the Navy.

Mr. Rydz received the Bachelor of Science degree in physics from MIT, Cambridge, and a Master of Science degree in physics from the University of Pennsylvania, Philadelphia. He is a member of the Optical Society of America and the Technical Association of the Graphic Arts.

❖

(Continued on page 54A)

IRE People

(Continued from page 52A)

The appointment of **Monroe H. Postman** (S'44–A'50–M'55) as Western Regional Sales Manager for Bryant Computer Products Division of Ex-Cell-O Corporation has been announced.

M. H. Postman

He will be headquartered in Sunnyvale, California.

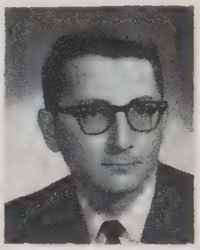
Prior to joining Bryant, he had been employed as Engineering Group Supervisor at the Philco Corporation Western Development Laboratory, Palo Alto, Calif.

He entered the electronics and computer fields in 1953, and since that time has worked in engineering capacities with the Monroe Calculating Machine Company, Bulova Watch Company, and New York University Univac Computing Facility.

He is a member of the IRE Professional Group on Electronic Computers.

❖

Motorola's Military Electronics Division has announced the appointment of **Arnold R. Sabel** (S'51–A'51–M'57) to its senior technical staff.



A. R. Sabel

In his new position, he will contribute to the development of long range technical plans for space and airborne electronics systems. His position is newly created, and reflects the company's increasing emphasis in this field of advanced aerospace systems engineering.

With Motorola for the past 10 years, his diversified engineering experience encompasses systems design and development, engineering management and tactical operations analysis and other systems studies. Most recently, he has served as Section Head of the Data Display Section of Motorola's Data Systems Laboratory. During 1957–58 he also served as member of an Office of Naval Research advisory committee with responsibilities for communications and data processing.

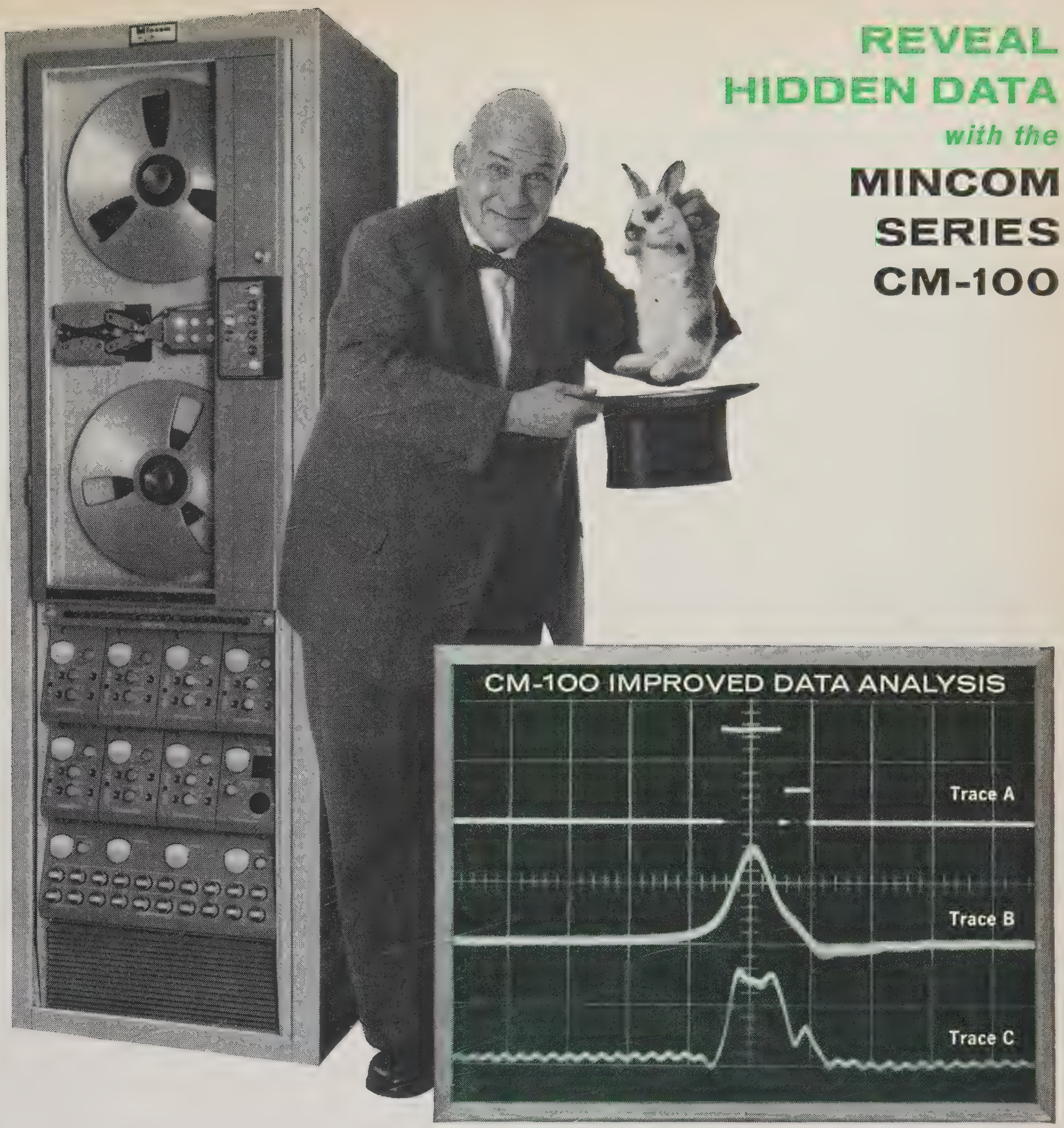
Sabel is a graduate of Purdue University, Lafayette, Ind., and is a member of Eta Kappa Nu.

❖

(Continued on page 56A)

Use your IRE DIRECTORY! It's valuable!

REVEAL HIDDEN DATA *with the* MINCOM SERIES CM-100



10 μs pulse separated from 4 μs pulse by 1.2 μs space. **Trace A:** 100-kc system input. **Trace B:** 100-kc output. **Trace C:** CM-100 output. Sweep Rate: 10 μs/cm. Vertical Deflection: .5v/cm.

Pulses recorded on any standard 100-kc system reveal previously undisclosed data when played back on the Mincom Series CM-100 Video Instrumentation Recorder/Reproducer. At 60 ips, a prerecorded tape from a standard 100-kc recorder will present on the CM-100 an improved frequency response of 200-220 kc $\pm$ 4 db with a practical limit of 250 kc. CM-100's superior playback heads and phase-compensating electronics produce better rise time, correcting for phase shift and overshoot. This recovery of hidden data is only one of the advantages of the CM-100, a 7 or 14-track 1-megacycle system which is now performing predetection recording/reproducing on an operational basis—in FM, FM/FM modulation, PCM and PCM/FM. Write for specifications.

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IRE People

(Continued from page 54A)

Lawrence J. Torn (M'50–SM'56) has been appointed general manager of the new Data Systems Division of Harman-Kardon, Inc.

L. J. Torn

A native of Fort Lauderdale, Fla., Mr. Torn comes to Harman-Kardon from Airborne Instruments Laboratory where he was general manager of the Apparatus Division. He received the B.S. degree in electrical engineering from Pennsylvania State University, University Park, and the M.E.E. degree from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. He is a member of Tau Beta Pi.

❖

Dr. H. C. A. van Duuren (SM'58–F'61) has received the De Groot Award in the Netherlands for his outstanding contributions to radio telegraphy.

H. van Duuren

In the Netherlands, Dr. van Duuren is well known as the inventor of the ARQ system, which provides for the automatic error correction of telegraphing signals by means of a return path. In developing this system, he also invented his own code converter for changing the normal 5-unit telex code into a 7-unit code which would allow error detection. The equipment as a whole is known as TOR (Telex on Radio), and many such units have been produced in the Netherlands PTT and elsewhere.

Dr. van Duuren graduated from the Technische Hogeschool at Delft in 1926. His entire career has been with the Netherlands Postal and Telecommunications Service, and since 1954 he has been the Director of the Dr. Neher Laboratories at Leidschendam near The Hague. He is a member of the Netherlands U.R.S.I. Committee and the Chairman of Study Group III (Fixed Service Systems) of the C.C.I.T.T.

The De Groot Award is given approximately once every five years and is in memory of Dr. C. J. de Groot, who established radio communication between Java and Holland during the First World War. His system employed a Poulsen arc transmitter and a large antenna slung across the sides of a deep valley. To obtain a true picture of the feat, one would have to know of the many improvisations that had to be made—such as using the motors from street-cars!

The year 1960 also witnessed the 40th anniversary of the Nederlands Radio-

(Continued on page 58A)


IS THERE
A FOREST
BEYOND
THE TREES?
The same can be asked of men and their works. In turn, a company is revealed by its products—the signs of a fertile, productive technology or remnants of a declining art.
Fairchild Semiconductor first revealed itself with two diffused silicon transistors. They were different. The technology advanced quickly to the production of silicon transistors that challenged the speed of germanium ... then yielded Planar transistors and diodes where an integral oxide surface achieved a new reliability. Planar in turn has led to practical Micrologic elements and the Planar Epitaxial transistor. Fairchild products have become the most copied in the industry.
If you value the satisfaction of working in a fast-paced technology and yours is a relevant background, we would like very much to hear from you.
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A wholly owned subsidiary of Fairchild Camera and Instrument Corporation

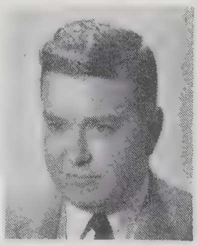
IRE People

(Continued from page 56A)

genootschap (Netherlands Radio Society), and the award was made to coincide with this event.

❖

Don White Associates, a technical consultant firm specializing in electronic systems analysis, has established private consulting offices in Bethesda, Md. Headed by **Donald R. J. White** (M'52–SM'56), it offers technical guidance and staff support on problems of radio-frequency interference prediction and control, mathematical modeling and evaluation of electronic weapons systems, operations analysis of communication-electronic systems, and electronic intelligence research programs.



D. R. J. White

Mr. White holds the B.S.E.E. and M.S.E.E. degrees from the University of Maryland, College Park, and has thirteen years of experience in directing electronic system design and evaluation programs, including radar, communications, navigation, ECM, and intercept and detection systems. He is a registered professional engineer, has served as Chairman of the Third National IRE Symposium on Radio-Frequency Interference, and is a member of Tau Beta Pi. He has a number of patents pending and is the author of more than twenty technical papers published in professional journals and technical periodicals.

❖

W. O. Swinyard (A'37–M'39–SM'43–F'45), a founder and former President, has been named Chairman of the National Electronics Conference for the third time. He is the first to be so honored by the Conference, which is an organization initiated in 1944 by a group of pioneers in electronics from education and industry, in order to "advance the science of electronics and its application and use in the pulbic interest and for the public good."

A Vice President and Director of Hazeltine Research, Inc., a subsidiary of Hazeltine Corporation, Mr. Swinyard is a Fellow of the American Association for the Advancement of Science and the Radio Club of America. He is also a former director of the IRE and past chairman of its Chicago section.

❖

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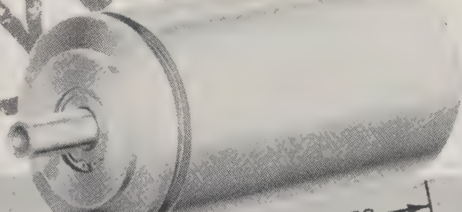
Oster® servos



Size 5 Type 5351-01

1" MAX. 0.7 OUNCES

This is a representative list. Many other servo types immediately available from stock.

SIZE	VOLTAGE (FREQUENCY 400 CY.)	STALLED TORQUE In. Oz. (MIN)	FP/CP IMPEDANCE AT STALL OHMS	NO LOAD SPEED RPM (MIN)	INERTIA gm.cm.²	THEORETICAL ACCELERATION AT STALL Rad./Sec.²	TEMP. RANGE	OSTER TYPE
5	26V	0.12	325/325	9,500	0.18	47,000	−55°C +125°C	5351-01
8	26V/26V	0.15	288/288	6,200	0.47	22,500	−54°C +125°C	5004-01
8	26V/36V CT	0.15	288/530	6,200	0.47	22,500	−54°C +125°C	5004-02
8	26V/36V CT	0.33	144/271	6,500	1.07	21,800	−54°C +125°C	5002-04
8	26V/40V CT	0.20	230/519	6,200	0.47	30,000	−55°C +125°C	5004-09
10	18V/18V	0.13	144/144	10,000	0.7	13,100	−55°C +71°C	5051-11
10	26V/26V CT	0.25	173/173	10,000	0.46	38,400	−54°C +85°C	5051-21
10	18V/54V	0.13	160/1470	10,000	0.46	19,500	−54°C +85°C	5058-02
10	18V/18V	0.13	160/160	10,000	0.46	19,500	−55°C +85°C	5051-26
10	26V/26V	0.30	157/157	6,500	0.46	46,000	−54°C +125°C	5052-60
10	27.5V/14V	0.15	1200/195	5,400	1.85	5,720	−54°C +85°C	5054-01
10	26V/35V	0.24	100/268	4,000	0.46	37,000	−55°C +85°C	5726-17
10	54V/54V	0.13	1470/1470	10,000	0.46	19,500	−55°C +125°C	5058-14
11	115V/180V	0.60	2175/5000	6,200	1.07	41,000	−54°C +85°C	5101-18
11	115V/115V	0.60	2175/2175	6,200	1.07	41,000	−55°C +125°C	5104-01
11	115V/57.5V	0.60	2175/2175	6,200	1.07	41,000	−55°C +125°C	5101-73
11	115V/40V/20	0.63	2175/2175	6,700	1.07	41,000	−54°C +120°C	5101-35
15	115V/115V	1.45	1030/1030	5,000	3.3	31,000	−54°C +75°C	5153-06
18	115V/300V	2.25	640/4500	5,000	4.0	39,700	−55°C +75°C	5201-01
18	115V/115V	2.25	640/640	5,000	4.0	39,700	−54°C +75°C	5201-08

SIZE 5 TYPE 5351-01 CAN ALSO BE FURNISHED WITH 18V OR 33V

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Engineers For Advanced Projects: Interesting varied work on designing transistor circuits and servo mechanisms. Contact Mr. Dallas Nielsen, Personnel Manager, in confidence.

Industrial Engineering Notes*

Association Activities

The Supplementary Report of the Television Allocations Study Organization has been published and can be ordered from Dr. G. R. Town, Dean, College of Engineering, Iowa State University, Ames, Iowa. Price of the report is $10 a copy. Checks should be made out to Television Allocations Study Organization.

FCC Actions

Long-awaited approval of the Hartford Phonevision Company's application for the first trial of subscription television has been granted by the Federal Communications Commission. The Hartford firm is licensee of TV station WHCT, a UHF outlet in Hartford, Conn., which will be used for three years of pay telecasting. The concern was the only applicant for such tests under conditions laid down by the FCC in 1959. Participating in the Hartford test will be RKO General, Inc., sole owner of Hartford Phonevision Company and local franchise holder for the "Phonevision" subscription system to be used in the trial; Zenith Radio Corporation, developer and patent holder of that system; and Television Entertainment Company, Inc. (Teco), Zenith's patent licensee. The text of the Commission's Public Notice follows: "RKO has assumed the expense of the Hartford test, which it estimates will cost up to $10 million and which it does not expect to recover during the three-year period. WHCT contemplates airing selected subscription programs, without commercials, about 40 hours a week, of which some 17 hours would be unduplicated. Approximately 30 additional hours will be conventional 'free' TV fare. It expects first-run films to predominate the pay programs, with stage, music, educational and sports as added attractions. The two types of programs will not be transmitted simultaneously. All of WHCT's broadcasting will be in monochrome. Zenith's Phonevision system employs an encoder (which scrambles both picture and sound transmission) and a decoder (descrambler) connected to the subscriber's set. The knob of the decoder is turned to an index number for a desired pay program and an automatic switch then shifts from free to fee broadcast. The program identification numbers are expected to be furnished the subscribers by newspaper publication or program booklet distribution. A credit-type of decoder is proposed initially. Program usage will be recorded on billing tape inside the decoder. The customer is expected to examine the tape (probably monthly), add up the charges, and send the tape and payment to the station. Subscribers will not be required to purchase decoding or other special equipment. The only charges contemplated, other than for programs, are for installation and rental of the decoder. The estimated installation charge is from $7.50 to $10.00. The maintenance charge, covering maintenance and repairs and depreciation, is expected to be from 75 cents a week to $28.00 a year. The per-program charge will range from about 25 cents to $3.50, depending upon cost factors, with the majority of programs being in the price range of 75 cents to $1.50. Charges would not vary with the number of viewers in a subscriber's home. If the same feature film is run twice in the same evening, plans call for one charge per set for both showings. Decoders will be installed only at locations which can receive WHCT's signal satisfactorily. Commencement of operations is planned after 2,000 decoders have been installed, which WHCT expects will require six months, with expectation of 10,000 installations at the end of the first year. The Commission's Report and Decision which made this grant after a hearing (Docket 13814) emphasizes its trial nature and subjects it to certain conditions. (Text to be printed by the Government Printing Office in weekly pamphlet.) The authorization is for the three-year period prescribed for such trials, subject to termination if WHCT's license should not be renewed during that time. WHCT's trial period will start with its first transmission of pay programs to subscribers."

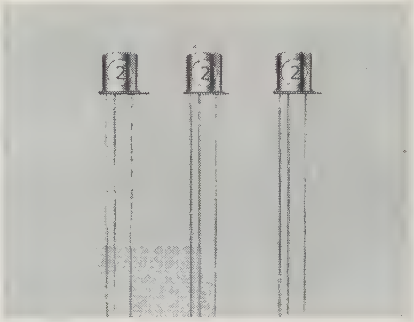
Governmental and Legislative

The Federal Aviation Agency's second year was highlighted by the expansion and continued progress toward the automation of its air traffic control system, according to the Agency's Second Annual Report. The report, covering calendar year 1960, showed completion of FAA traffic control center buildings at Oakland, Atlanta, Cleveland, and Jacksonville. Construction on new buildings also was started for the Indianapolis, Kansas City, Fort Worth, Chicago, Salt Lake City, Seattle, Memphis, Minneapolis, and Denver air route traffic control centers. In addition, sites for new center buildings were selected for the Boston, New Orleans, Washington, D. C., Los Angeles, New York City, and Miami centers. Calendar year 1960 also saw the delivery of the first components of a new semi-automatic air traffic system called the Data Processing Central. This computer and its associated equipment will print flight progress strips, update and memorize information, exchange such information

(Continued on page 62A)

* The data on which these Notes are based were selected by permission from *Weekly Report* issues of February 27 and March 6, and 13, 1961, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.

Three New Additions to the Sprague MADT* Transistor Line



The Sprague Electric Company has added a new series to their highly-successful line of Micro-Alloy Diffused-base Transistors.

The new units, Type 2N768, 2N769, and 2N779A, are high-speed switching transistors in TO-18 cases. Their unique electrical characteristics further expand the varied applications to which Sprague MADT Transistors can solve circuit design problems.

Type 2N768 is a micro-energy switch designed for low current, low voltage, high speed applications.

Type 2N769 is the fastest switching transistor yet developed. It will operate reliably at speeds in excess of 100 mc.

Type 2N779A is manufactured with tighter parameter control than any other transistor in the industry. It is ideally suited for NOR logic and other super-critical applications.

These hermetically-sealed germanium transistors are made by a controlled-etch process to insure extreme uniformity. Maximum frequency capabilities have been improved by graded-base construction. Automated manufacturing techniques have brought about increased production efficiency, permitting favorable reductions in prices. This is why Sprague MADT Transistors can offer you greater performance per dollar than other high-speed devices in low-current switching circuits.

For prompt application engineering assistance, write Commercial Engineering Section, Transistor Division, Sprague Electric Company, Concord, N.H.

For complete engineering data sheets, write Technical Literature Section, Sprague Electric Company, 235 Marshall St., North Adams, Mass.

*trademark of Philco Corp.

Industrial Engineering Notes

(Continued from page 60A)

with other traffic control facilities and warn the controller of possible conflicting flight plans. The first of these computers is scheduled to be installed at the Boston traffic control center in October, 1962. Also during the past year, the FAA put into effect several new major Civil Air Regulations. One required that air transports be equipped with airborne weather radar. Another specified the installation of flight recorders on all turbine powered transports. . . . **Investigations of space communications policy, use of the radio spectrum, television frequency allocations, pay TV, and community antenna and booster problems have been slated by the Senate Interstate and Foreign Commerce Committee.** The committee's agenda, outlined in a Senate-approved request for $315,000 for this year's operations, showed projected studies of a policy of space communications by satellites and a review of military and civilian use of the radio spectrum. In addition, the committee will consider legislation dealing with common carriers in domestic and international communications. Allocation of TV channels will also be studied with the objective of establishing a "nationwide competitive television service." The committee said it will also examine subscription and education TV and the "extremely controversial subject" of community antennas and boosters. The committee is chaired by Sen. W. G. Magnuson (D. Wash.). . . . **The Federal Communications Commission's inquiry into space communications brought four major proposals for the use of satellites to ease congestion of earth-bound communications systems and make possible world-wide bouncing of radio signals.** The General Electric Company, in its filing in the FCC's Docket 13522, envisions a 10-satellite, active relay system. The satellites would be maintained in equatorial orbits 6000 miles high in positions fixed to one another and to points on the earth's surface. The Radio Corporation of America's proposed system involves three satellites in equatorial orbit at 22,000 miles for world-wide coverage. Ground stations would control both the position and altitude of the satellites. They would orbit at the same speed as the earth's rotation, thus maintaining a fixed position in space. The American Telephone and Telegraph Corporation recommends a 50-satellite system with active relay units at heights of 2000 to 8000 miles. AT&T favors active repeater satellites because of their low power consumption. Lockheed Aircraft Corporation proposes a joint common carrier satellite which the major communications firms would finance jointly. Under the Lockheed plan, common carriers would lease communications facilities for world-wide services.

(Continued on page 64A)

Industrial Engineering Notes

(Continued from page 62A)

Industry Marketing Data

Japanese electronics production during the first nine months of 1960 totaled $856 million, a 31 per cent increase over the $655 million total during the corresponding period 1959, the Electronics Division, Business and Defense Services Administration, U. S. Department of Commerce, has reported. Production of television receivers and radio receivers accounted for one-half the total Japanese electronics output in the first nine months 1960. Production of these items which expanded rapidly in 1959 showed signs of leveling off during the third quarter 1960, while a strong upward trend appeared in the production of tape recorders and radio phonographs. Gains in production of these items in the third quarter 1960 over third quarter 1959 output were as follows: Television receivers, 2 per cent; radio receivers having 3 or more transistors, 21 per cent; radio-phonographs, 134 per cent; and recorders, 81 per cent. Output of tube type radios declined 29 per cent in the third quarter 1960 from the third quarter 1959. Significant gains were made in Japanese production of a number of other products in January–September, 1960 over January–September, 1959. Electronic computers increased by 172 per cent, industrial measuring and control equipment, 73 per cent; receiving tubes, 45 per cent; transmitting and special purpose tubes, 50 per cent; transistors, 34 per cent; transformers, 87 per cent; and amplifiers for "hi-fi," 157 per cent, to $3.2 million. . . . **Rapid expansion of production in the computing machines industry has increased industry employment by more than 50 per cent over the past five years, according to a survey released recently by the U. S. Labor Department's Bureau of Employment Security.** The Bureau said computing machines industry employment increased from 64,700 in 1956 to 103,000 in September, 1960. The survey covered 68 firms producing computing and accounting machines, including cash registers. These firms accounted in September for 95 per cent of employment in the computing machines industry. The Bureau said the outlook for the computing machines industry was for continued expansion in employment and production. It pointed out that while the use of computers to mechanize clerical and bookkeeping functions and for scientific and engineering calculations has been rapid and impressive, the greatest growth potentiality lies in the future, particularly in the application of computers to control production processes. . . . **United States imports of electronic products increased to $103 million during 1960, or a total 25 per cent higher than that for 1959, according to a report by the Electronics Division of the Business and Defense Services Administration.** The increase, BDSA said, is notably lower than the rate for the first six months of the past year, when imports were about 60 per cent greater than in the same 1959 period. Principal factors accounting for the trend were the sharp increase in imports of radio apparatus and parts in the latter half of 1959 and a leveling off of these imports in 1960. Of total imports of this group of items, transistor radios, mainly from Japan, accounted for about 60 per cent, according to the report. In mid-1960, Japan imposed quotas on exports of radios containing three or more transistors, thereby preventing further sharp increases in exports to the U. S. U. S. exports of electronic products in 1960 were estimated in the BDSA report to have reached a record high of more than $450 million, more than 12 per cent over 1959. Large gains were made by electronic detection and navigation apparatus, semiconductor devices, electron tubes, electronic computers, and electronic test equipment. BDSA noted that U. S. electronic products were shipped to practically every country in the free world, although in the last two years, 15 countries accounted for over 80 per cent of the total value of electronics exports, excluding "special category" items, for which countries of destination were not available. Canada was by far the largest single market, BDSA said, but other important markets were Mexico, Venezuela, Argentina, West Germany, Italy, France, the United Kingdom, and Japan. The loss of the substantial Cuban market and some loss in Venezuela have been offset by gains elsewhere, particularly in West Germany, Italy, and Japan.

Military and Space

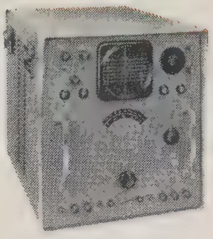
The emphasis in military, solid state electronics in the future will not be on the development of equivalent or better replacements for present conventional electronic components, but in a basic change in system design, circuit design, and devices fabrication, according to the final report of an Office of Naval Research study group published by the Commerce Department. . . . A survey of research and development work at 22 private organizations and several government laboratories conducted by the ONR study group indicates a bright future for military microelectronics with increasing chances for the achievement of planned objectives. A list of component size improvements achieved as well as anticipated, which was established as a result of the survey, is included. The report is *Final Report of ONR Study Group on Microelectronics.* A. Brodzinsky, Office of Naval Research, June, 1960. 20 pages. (Order PB 161 890 from OTS, U. S. Department of Commerce, Washington 25, D. C., 50 cents.) . . . **A new Aerospace Surveillance and Control Squadron which will detect, identify and track man-made objects in space has been established by the Air Defense Command at Colorado Springs, Colo.** According to a Defense Department announcement, the squadron will supply data on space objects "of potential or actual military concern" to the North American Air Defense Command. DOD also announced that unclassified scientific orbital and trajectory information on space previously published by the National Space Surveillance Control Center, Air Research and Development Command, henceforth will be handled by the National Aeronautics and Space Administration, Goddard Space Flight Center, Greenbelt, Md.

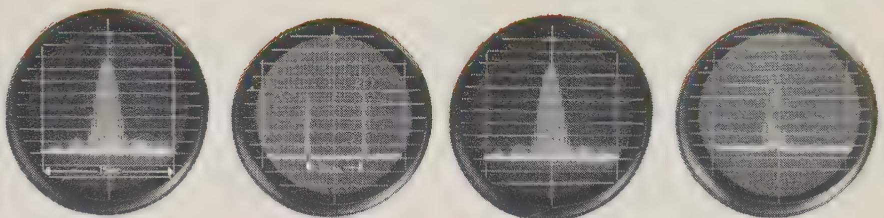
POLARAD

VISUAL MICROWAVE ANALYSIS 10 to 44,000 mc

MODEL TSA
DIRECT-READING SPECTRUM ANALYZER

10 to 44,000 mc with five plug-in tuning units

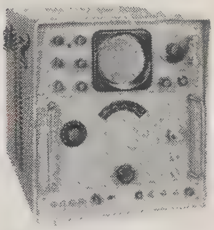


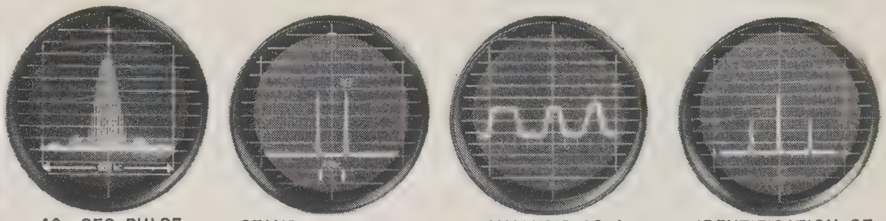


0.4 μSEC PULSE | STANDARD SIGNAL | DECODED MULTIPULSE SPECTRUM | INCIDENTAL FM ANALYSIS

MODEL TSA-S
COMBINATION SYNCHROSCOPE-SPECTRUM ANALYZER

10 to 44,000 mc with five plug-in tuning units

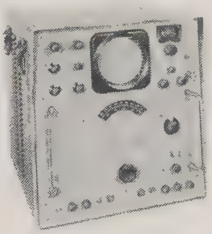


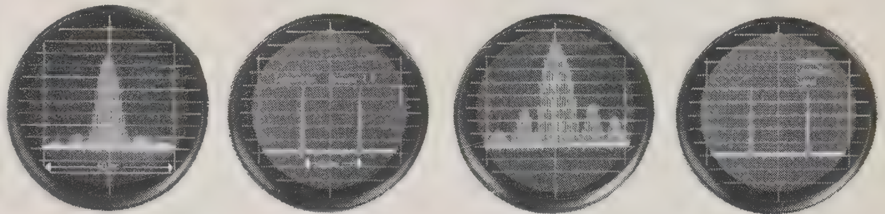


10 μSEC PULSE | STANDARD SIGNAL | ANALYSIS AS A FUNCTION OF TIME | IDENTIFICATION OF AMPLITUDE MODULATION

MODEL TSA-W
WIDE DISPERSION SPECTRUM ANALYZER

10 to 44,000 mc with five plug-in tuning units—70 mc dispersion

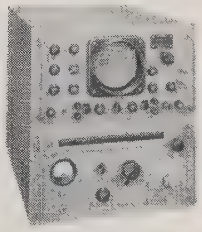


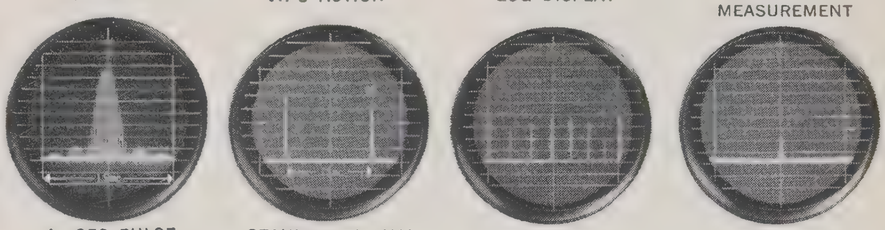


0.1 μSEC PULSE | AFC ACTION | LOG DISPLAY | SIGNAL STABILITY MEASUREMENT

MODEL SA-84
UNIVERSAL SPECTRUM ANALYZER

10 to 44,000 mc in one integrated self-contained unit

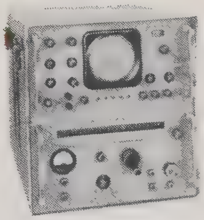


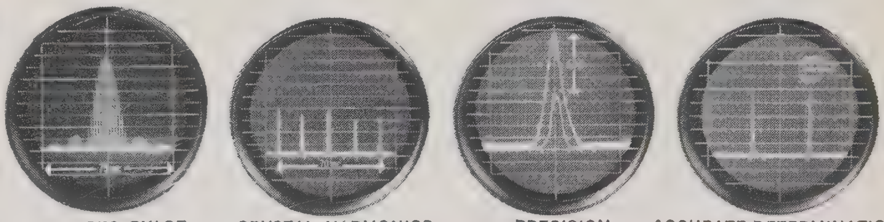


4 μSEC PULSE | STANDARD SIGNAL | FM SIGNAL ANALYSIS | LEAKAGE AND RADIATION MEASUREMENT

MODEL SA-84W
WIDE DISPERSION UNIVERSAL SPECTRUM ANALYZER

10 to 44,000 mc in one integrated self-contained unit—features over 80 mc dispersion





0.08 μSEC PULSE | CRYSTAL HARMONICS | PRECISION ATTENUATION | ACCURATE DETERMINATION OF FREQUENCY

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Section Meetings

Akron

"Achieving a Reliable System," G. Raymond, Remington Rand; Presentation of Fellow Award. 1/17/61.

"Signal Processing in the TIROS Instrumentation System," M. H. Mesner, RCA Labs. 2/21/61.

Student Night. "Transistorized Flip-Flops," J. Cozzens; "Industrial Process Controls," J. Van Oss; Tour of Univ. of Akron, EE Labs. 3/14/61.

Alamogordo-Holloman

"Advancing Technology in the Space Race," R. K. Sherburne, U. S. Army. 2/21/61.

Anchorage

"Parametric Amplifiers," D. Bryce, Univ. of Alaska. 1/6/61.

Atlanta

"Transistors," W. F. Tewksbury, Bend x Radio Div. 2/24/61.

Baltimore

"Science in Crime Detection," I. Conrad, FBI Labs. 2/6/61.

"Navigation by Means of Artificial Satellites," G. C. Weiffenbach, Johns Hopkins Univ. 3/13/61.

Bay of Quinte

"Binaural Systems As an Aid to Teaching the Deaf," J. Boyd, Ontario School for the Deaf, and A. Jamroz, Northern Electric Co. 2/22/61.

Beaumont-Port Arthur

"Five Points for National Defense," B. Schiever, USAF. 2/23/61.

"Transmitter and Antenna Systems," M. Bulloch, Continental Electronics. 3/7/61.

Benelux

"Automation of Accounting and Arithmetical Processes," J. M. Unk, Philips Telecommunications Industry. Presentation of Fellow Awards. 2/25/61.

Binghamton

Technical Description and tour of WNBF-TV-FM-AM, L. Cox, WNBF. 2/15/61.

Central Florida

"Who Should Manage Engineers," Panel discussion with Mr. Warren, G. E.; N. Chase, Lockheed; Mr. McNabb, Convair; Mr. Kimball, Minute-Maid; Mr. Smith, Orlando Utilities. 1/30/61.

"Satellites and Communications," G. S. Shaw, Radiation, Inc. 2/16/61.

Central Pennsylvania

"History and Development of Nuvistors," J. Halgren, RCA; "A Developmental Nuvistor Triode for Small Signal Grounded Grid Amplifiers," R. J. Rundstedt, RCA. 2/21/61.

Chicago

"Russian Management and Technology," E. L. Ryerson, Inland Steel Co. 2/10/61.

"The Research and Development Activity at Teletype," W. J. Zenner, Teletype Corp. 3/2/61.

China Lake

"Detection and Measurement of Infrared Radiation," D. E. Martz, USNOTS, China Lake. 3/6/61.

Cincinnati

"AVCO Automatic Data Communication System," E. W. Moulton, AVCO Electronics Div.; "Recent European Audio Research," D. W. Martin, Baldwin Piano Co. 2/21/61.

Colombia

"Electronics in the Artificial Heart-Lung," A. Vejarano and J. Reynolds, Shaio Foundation. 2/17/61.

Columbus

"Use of Satellites for Communications," G. E. Mueller, Space Technology Lab. 2/14/61.

"An Electronic Reader for the Blind," J. S. Abma and L. Mason, Battelle Memorial Institute, 3/8/61.

Dallas

"Measurement and Analysis of Magnetotelluric Signals," H. W. Smith, Univ. of Texas. 2/23/61.

"Human and Biological Factors in Space," H. G. Clamann, Brooks AFB. 3/2/61.

"Planet Exploration," E. Stuhlinger, NASA. 3/9/61.

Detroit

"Single-Sideband Engineering," W. B. Bruene, Collins Radio Co. Presentation of Fellow Awards. 2/17/61.

Egypt

"Memory Devices—Introduction and Examples of Low-Speed, High-Capacity and High-Speed Low-Capacity Memory," A. L. Ahmed, UAR Broadcasting & TV. 3/4/61.

Elmira-Corning

"Transparent Phosphors," C. Feldman, Melpar, Inc. 1/16/61.

"Ceramic and Ceramic-to-Metal Seals for Electronic Devices," M. Berg, RCA, 2/20/61.

Emporium

"Electronic Problems in Underwater Acoustics," W. Houser, Pennsylvania State University. 2/21/61.

Florida West Coast

"Ballistic Missile Early Warning System," H. W. Phillips, RCA-BMEWS. 2/15/61.

"Electronic Scanning Antennas," R. K. Thomas, The Martin Co., 3/15/61.

Fort Huachuca

"Speech Transmission by Selective Amplitude Sampling," L. R. Spogen, Fredericks Research Associates. 2/27/61.

Fort Worth

"The New York UHF Project," E. W. Allen, Federal Communications Commission. 2/14/61.

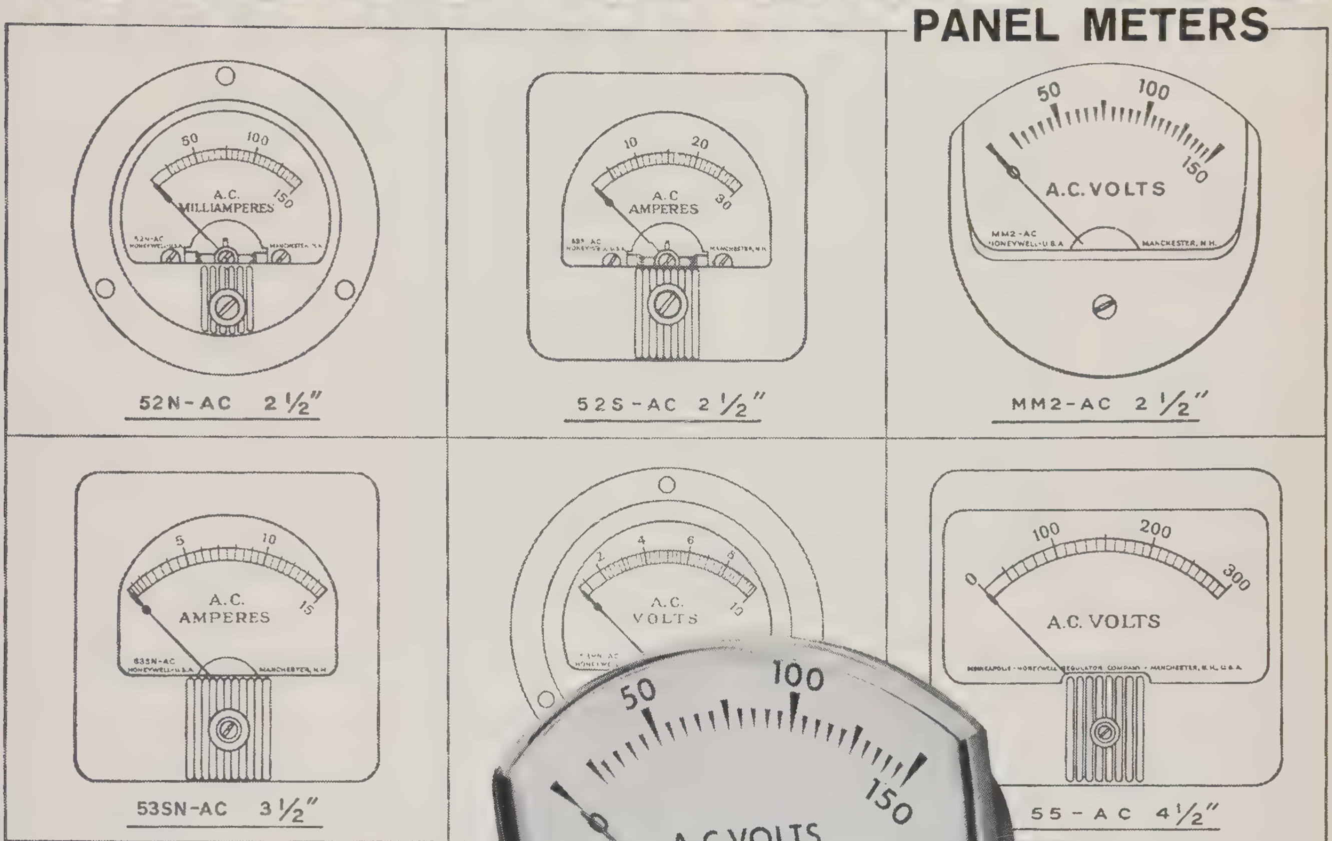
"Ferrites as Circuit Elements for Electrically Controlled Microwave Devices," W. H. von Aulock, Bell Telephone Labs., Inc. 3/14/61.

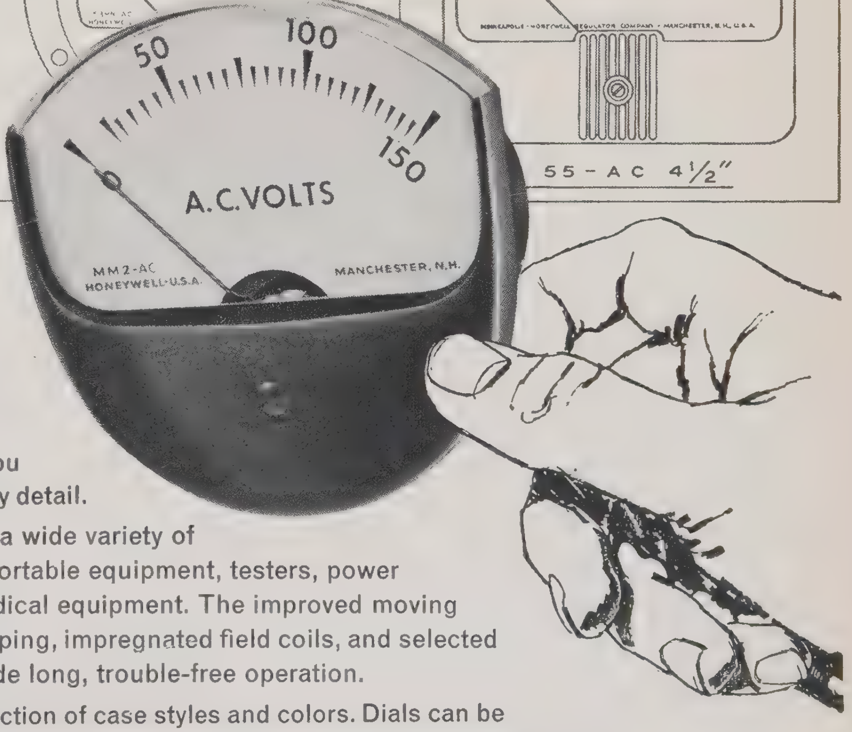
(Continued on page 68A)

ANNOUNCING THE NEW HONEYWELL

IRON VANE AC

PANEL METERS





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Iron Vane AC Meters are designed for a wide variety of commercial applications — including portable equipment, testers, power supplies, generator equipment and medical equipment. The improved moving iron mechanism features magnetic damping, impregnated field coils, and selected fixed and moving iron material to provide long, trouble-free operation.

These meters are available in a wide selection of case styles and colors. Dials can be custom designed with your company name, trade-mark or other data. For full information, contact our representative in your area — he's listed in your classified telephone directory. Or us: Precision Meter Division, Minneapolis-Honeywell Regulator Co., Manchester, N.H., U.S.A. In Canada, Honeywell Controls Limited, Toronto 17, Ontario and around the world: HONEYWELL INTERNATIONAL — Sales and service offices in all principal cities of the world.

Honeywell

H Precision Meters

Section Meetings

(Continued from page 66A)

Gainesville

"Parametric Amplifiers," A. D. Sutherland, Sperry Rand Corp. 2/8/61.

"Superconductivity and Its Applications," T. A. Scott, Univ. of Florida. 3/8/61.

Geneva

"Stereophonic Broadcasting Systems," F. L. Stumpers, Philips Research Labs. Election of Officers. 2/27/61.

Huntsville

"Design Parameters for Missile Transducers," R. J. Horak, Consolidated Electro Dynamic Corp. 1/26/61.

"Plasma Characteristics and Uses," C. Cason, Army Rocket & Guided Missile Agency. 2/23/61.

Kansas City

"PERT—Program Evaluation and Review Technique," L. F. Gardner, U. S. Navy. 2/21/61.

"Electric Spaceship Propulsion," M. Ghai, G. E. Propulsion Lab. 3/14/61.

Kitchener-Waterloo

"Some Problems in the Development and Production of Antennas," W. V. Tilston, Sinclair Radio Labs., Ltd. 2/20/61.

Little Rock

"Communication System for Project Mercury Space Capsule," W. Benner, McDonnell Aircraft Corp. 2/13/61.

London (Ontario, Canada)

"Research in Semi-Conductors," J. M. Stewart, RCA Victor Co., Ltd., 2/20/61.

Long Island

"Stereo-Manship and Such," B. B. Bauer, CBS Labs. 1/10/61.

"Experiments in Machine Learning Using the Game of Checkers," A. Samuel, IBM. 2/14/61.

Los Angeles

"Government and Science—The Changing Pattern," C. Sherwin, Aerospace Corp.; "Progress Report—Harvey Mudd College," J. Platt, Harvey Mudd College. 2/7/61.

Lubbock

"Radio Activity and the Effects of Nuclear Weapons," R. W. Marsh, Reese AFB. 2/28/61.

Milwaukee

"Elementary Particles," C. E. Falk, Brookhaven National Laboratory. 3/7/61.

Montreal

"The Implementation of an Upper Atmosphere Chemical Seeding Experiment," G. B. Spindler, C.A.R.D.E. 2/13/61.

"Electronics as Applied to Life Insurance," S. J. Chapman, Sun Life Assurance Co. of Canada. Visit to UNIVAC II Computer. 2/22/61.

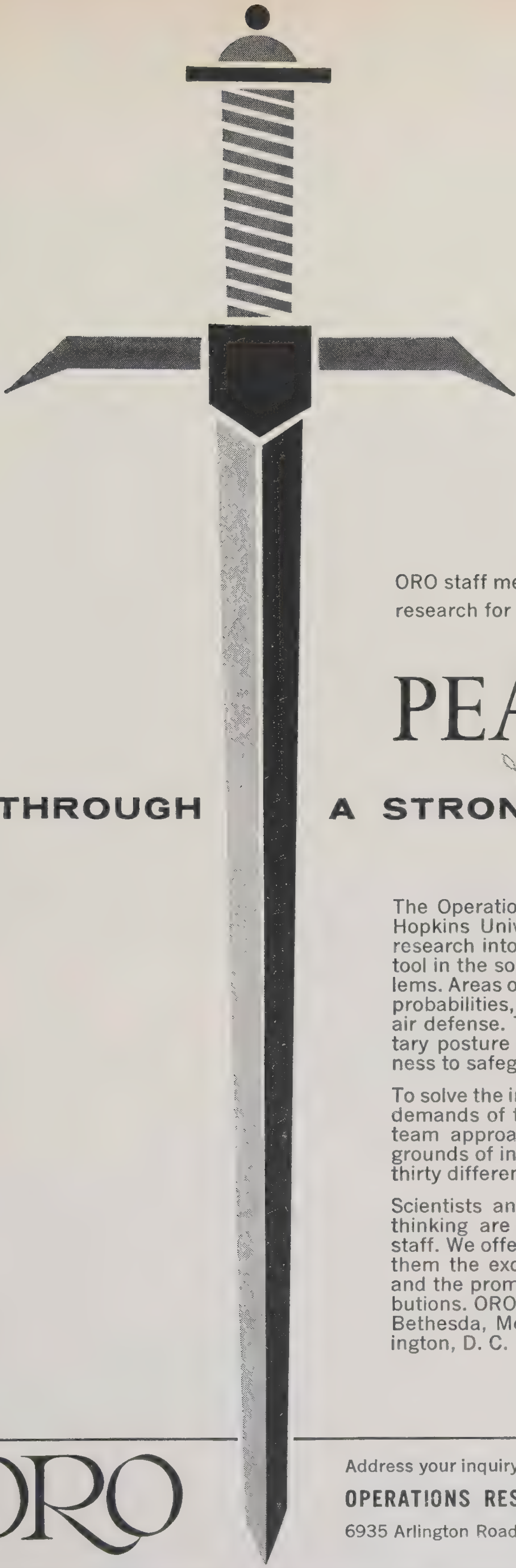
New York

"The Thin-Film Cryotron," R. B. DeLano, IBM Corp. 3/1/61.

North Carolina

"The Ballistic Missile Early Warning System (BMEWS)," G. A. Thorpe, Western Electric Co. 2/17/61.

(Continued on page 70A)



Section Meetings

(Continued from page 68A)

NORTHERN ALBERTA

"A New Concept in Point-to-Point Radio Systems for Multi-Channel Service," W. C. Fisher, Farinon Electric. 2/15/61.

NORTHERN NEW JERSEY

"Should Engineers Do Their Own Writing?" a debate by J. D. Chapline, Philco Corp., and T. R. O'Neill, Cornell Design Co. 2/8/61.

"Electronics Applied to Home Appliances," G. W. Schroeder, G.E. Co. Presentation of Fellow Awards and Special Awards. 3/8/61.

PHILADELPHIA

Annual Fellow Awards dinner and presentation of Philadelphia Section Award. 2/11/61.

PITTSBURGH

"Extreme Adaptation in the Presence of Noise," R. Van Nice, Westinghouse Research Labs. 1/26/61.

"The Optical Maser," R. D. Haun, Jr., Westinghouse Research Labs. 2/6/61.

"Electric Home of the Future," F. Maleck, Westinghouse Electric Corp. 3/6/61.

QUEBEC

"Microwave Components for the 1.5–3.0 Millimeter Wave Region," L. L. Bertan, FXR, Inc. 2/28/61.

RIO DE JANEIRO

"Recent Progress in the Technology of Millimeter Waves," L. C. Bahiana, 11/9/60.

"Electronics and Nuclear Energy," L. G. Langsch D., Brazilian Navy. 3/8/61.

ROCHESTER

"Solid State Logic, Control and Character Recognition Features of the Burroughs B100 Magnetic Check Sorter-Reader," T. A. Dowds and H. Rosenberg, Burroughs Corp. 2/16/61.

SACRAMENTO

"Time—Slow or Fast," I. J. Sandorf, Univ. of Nevada. 2/23/61.

SAINT LOUIS

"The Present and Future Status of Launch Vehicle Systems," D. R. Ostrander, NASA. 2/7/61.

"Communications Problems in the Surveyor Lunar Probe," Mr. Holwyk, McDonnell Aircraft Corp. 2/28/61.

SAN ANTONIO-AUSTIN

"Development of Ground Instrumentation for the Transit 1-B Satellite," G. G. Moore, Univ. of Texas. 11/17/60.

"Stereo in Home High-Fidelity Music Systems," P. W. Klipsch, Klipsch & Associates. 12/7/60.

"Electronic Data Processing," G. Countrymann, IBM Corp. 1/13/61.

SAN DIEGO

"Three-Level Ku Band Maser," E. G. K. Schwarz, General Dynamics/Electronics. 3/1/61.

SEATTLE

"Why Pulse Code Modulation," L. J. Spieker, Texas Instruments. 11/17/60.

"Getting Started in the Electronics Business," J. V. Granger, Granger Associates. 12/15/60.

"Microwave Tubes Come of Age," P. Crapuchettes, Litton Industries. 1/19/61.

"Engineering Management," panel discussion. Panelists: J. Fluke, J. Fluke Engineering Co.; W. R. Thompson, Boeing Airplane Co.; H. Vollum, Tektronix, Inc.; R. R. Dilling, Tally Register Corp.; F. Little, Boeing Airplane Co. 2/16/61.

SOUTH BEND-MISHAWAKA

"Radio Astronomy," G. R. Marner, Collins Radio Co. 12/1/60.

"Tetrode Power Transistors," J. Maupin, Minneapolis-Honeywell. 1/5/61.

"Application of Digital Techniques in Measuring Devices, Instrumentation and Control Systems," N. Kindt, Bendix Mishawaka Div.; "A Proposal to Standardize Power Supply Definitions," D. Dart, Bendix; "The Engineering Report—A Rusty Tool," R. Woelfle, Bendix; "Microwave Phase Locked Signal Generator Using Single Sideband Techniques," F. J. Schmiedeler and G. Julius, Bendix. 1/26/61.

"Thermo-electrically Controlled Temperature Chamber," A. Martz, Whirlpool Corp. 2/23/61.

TORONTO

"A Citizen's Life in Thule, Greenland," B. Davies, Canadian Westinghouse Co. 3/13/61.

VANCOUVER

Annual Students Night. Prizes in order: "A 20 mc/s Scale of 10 & a 20 mc/s Burst-pulse Generator," D. R. Heywood; "A Control System for a Heavy-Water Moderated Reactor," J. T. Walton; "Single Sideband Communications," B. E. Hanson; "Transistor Power Amplifiers," W. J. Harris. 2/20/61.

(Continued on page 72A)

Section Meetings

(Continued from page 70A)

WASHINGTON

"Digital Television," W. C. Coombs, National Bureau of Standards. 2/13/61.

WESTERN MASSACHUSETTS

"Professional Development," Miss Kellog, G.E. 1/24/61.

"Broad Application of Superconductivity," P. Schock, G.E. Co. 3/2/61.

SUBSECTIONS

BUENAVENTURA

"Parametric Amplifiers, Advantages, Limitations and Applications," A. D. Berk, Micromega Corp. 1/11/61.

"The Growth of WESCON and The Electronic Industry," W. Peterson, Automation Development Corp. 2/8/61.

BURLINGTON

Business meeting and election of officers. 3/2/61.

EAST BAY

"The Component Approach to Reliability," J. Lavrischeff, Univ. of Calif. 1/23/61.

EASTERN NORTH CAROLINA

"Operations at Oak Ridge," M. H. Lietzke, Oak Ridge National Laboratory. 2/10/61.

LANCASTER

"Photovoltaic Effects," J. J. Wysocki, RCA Labs. 1/24/61.

MEMPHIS

"Central Monitoring Facilities," H. Hanish, Dallons Labs., Inc. Election of officers. 1/27/61.

"Force-Reflecting Servomechanisms," R. Arzbaecher, Christian Brothers College. 2/24/61.

MID-HUDSON

"An Engineer Tackles Some Human Relations Problems," H. Rush, Opinion Research. 2/21/61.

MONMOUTH

"Peculiar Inventions," E. K. Kaprelian, USAS R & D Lab. Presentation of Fellow Award. 2/15/61.

NASHVILLE

"The Transistor as a Circuit Element," J. Wentworth, RCA. 3/8/61. Election of officers.

NEW HAMPSHIRE

"Instrumentation Utilizing Magnetic Tape Recording Techniques," J. Saret, Ampex Corp. 2/21/61.

ORANGE BELT

"History of Harvey Mudd College," J. Platt, Harvey Mudd College; "The Role of Government and Industry in Modern Technology," C. Sherman, Aerospace Corp. 2/7/61.

RICHLAND

"Instrument and Electrical Development Activities at ARCO," R. Crosier, Phillips Petroleum, ARCO. 2/15/61.

SAN FERNANDO VALLEY

Tour of NBC-TV Studios, Burbank, Calif. 2/22/61.

WESTERN NORTH CAROLINA

"Aleutian Skywatch" (Color Film), F. Houtz, Western Electric Co. 2/24/61.

- 10-25,000 TOTAL Q RANGE
- SELF-CORRECTING UHF RESONATING CAPACITOR
- DIRECT-READING INDUCTANCE SCALE
- 25 MV RF MEASURING LEVEL
- MEASURES "IN-CIRCUIT" Q OF SELF-RESONANT CIRCUITS

-610 MC

by BRC

Specifications

Radio Frequency Characteristics

RF RANGE: 210 to 610 MC
RF ACCURACY: ±3%
RF CALIBRATION: Increments of approximately 1%
RF MONITOR OUTPUT: 10 mv. minimum into 50 ohms*
*at frequency monitoring jack

Q Measurement Characteristics

Q RANGE:
Total Range: 10 to 25,000*
High Range: 200 to 25,000*
Low Range: 10 to 200
*10 to approx. 2,000 employing internal resonating capacitor

Q ACCURACY: ±20% of indicated Q

Q CALIBRATION:
High Q Scale: Increments of 1—5% up to 2,000
Low Q Scale: Increments of 3—5%

Inductance Measurement Characteristics

L RANGE: 2.5 to 146 mμh*
*actual range depends upon measuring frequency
L ACCURACY: ±11 to 15%*
*accuracy depends upon resonating capacitance
L CALIBRATION: Increments of approx. 5%

Resonating Capacitor Characteristics

CAPACITOR RANGE: 4 to 25 μμf
CAPACITOR ACCURACY: ± (5% + 0.2 μμf)
CAPACITOR CALIBRATION:
0.05 μμf increments, 4-5 μμf
0.1 μμf increments, 5-15 μμf
0.2 μμf increments, 15-25 μμf

Measurement Voltage Level

RF LEVELS: 25, 40, 80, 140, 250 mv. nominal*
*across measuring terminals

Physical Characteristics

MOUNTING: Cabinet for bench use; by removal of end covers, suitable for 19" rack mounting.

FINISH: Gray wrinkle, engraved panel (other finishes available on special order).

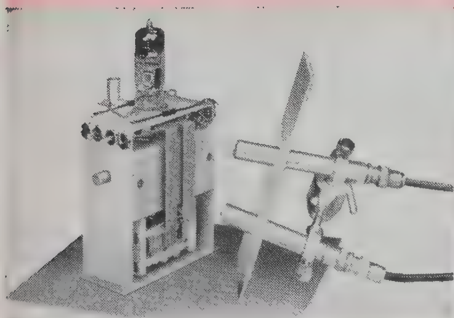
DIMENSIONS: Height: 12-7/32" Width: 19" Depth: 17"

WEIGHT: Net: 72 lbs.

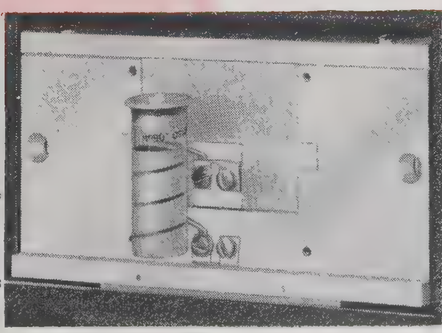
Power Requirements

280-A : 105-125/210-250 volts, 60 cps, 140 watts
280-AP: 105-125/210-250 volts, 50 cps, 140 watts

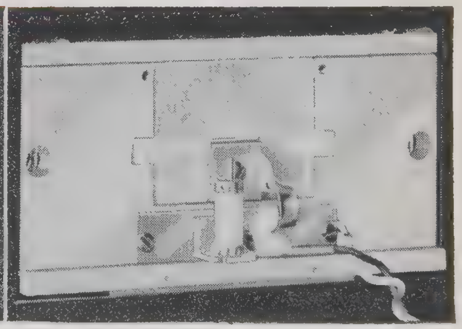
Price: 280-A: $2,375.00 280-AP: $2,375.00
F.O.B. Boonton, N. J.



"IN-CIRCUIT" Q MEASUREMENT



COIL MEASUREMENT



DIODE MEASUREMENT

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Professional Group Meetings

Aerospace and Navigational Electronics

Metropolitan New York—February 9

"Aerospace Application of Optical Masers," Dr. C. Ellis, General Precision, Inc.

Metropolitan New York—December 8

"Electrical Associated with Lunar Landings and Space Rendezvous," K. Stehling.

Metropolitan New York—November 10

"The Third Dimension in Air Traffic Control Radar," T. J. Simpson and K. F. Bierach.

Metropolitan New York—October 13

"Recent Developments in Small Boat Navigation Aids," L. E. Brunner, U. S. Coast Guard.

Aerospace and Navigational Electronics/Communications Systems

Florida West Coast—February 22

"Fluid Amplification," R. Warren, Diamond Ordnance Fuze Labs.

Antennas and Propagation

Chicago—December 9

"Atmospheric Transmission of Infrared Radiation," J. A. Sanderson, U. S. Naval Res. Lab., Washington, D. C.

Audio

Albuquerque-Los Alamos—February 15

"A Look at Tape Recorders," J. Mayfield, Audio Center, Albuquerque.

A question and answer session and demonstration of various tape recorders were conducted. A discussion of activities planned for the coming year was held.

Chicago—January 13

"An Improvement in Three-Channel Stereo," P. Tappan, Warwick Mfg. Co., Chicago, Ill.

San Francisco—February 15

"Advancements in Optical Sound Recording," W. A. Palmer, W. A. Palmer Films, Inc., San Francisco.

San Francisco—January 11

"The Vega Wireless FM Microphone," R. Tinkham, R. Z. Langevin, and R. Brown, Vega Electronics Corp., Cupertino, Calif.

Audio-Broadcasting

Washington—February 21

"Field Test of FM Stereo Multiplex, Report of Panel 5 NSRC," A. P. Walker, National Association of Broadcasters.

Automatic Control

Boston—February 16,

"Recoverable Space Probes," M. B. Tragerer, MIT Instrumentation Lab., Cambridge.

Los Angeles—February 14

"Astrodynamics as it is Related to Guidance and Control," R. M. L. Baker, Jr., USAF.

Bio-Medical Electronics

Buffalo-Niagara—January 17

"Electronic Aspects of Brain Tumor Localization," Dr. J. R. Hughes, Meyer Memorial Hospital.

Buffalo-Niagara—December 13

Panel Discussion—"Problems of AC Recordings from the Central Nervous System," Drs. W. H. Gillen, S. Wolering and W. Noell, University of Buffalo Medical School.

Cleveland—January 25

"Muscle—an Anatomist's View of Structure and Function," Dr. C. N. Loeser, Western Reserve University Medical School, Cleveland.

"Muscle—Thermodynamics and Molecular Aspects," W. Abrahamson, Case Institute of Technology.

Dr. A. Hopkins outlined the purpose of the Cleveland PGBME Section in presenting medical and engineering aspects of the basic divisions of medical science.

Cleveland—December 7

"Feedback Mechanisms in Reflex Activity of the Nervous System," Dr. W. F. Collins, Western Reserve University Medical School, Cleveland, Ohio.

"Electronic Feedback Systems," Dr. H. F. Klock, Case Institute of Technology, Cleveland, Ohio.

San Francisco—February 15

"Physiological Requirements and Problems Associated with Flight Motion Simulators and Airborne Vehicles," Capt. H. A. Smedal, USN(MC), Ames Res. Ctr., NASA, Moffett Field, Calif.

"Ames Physiological Instrument Systems," G. R. Holden, Ames Res. Ctr.

"Differential Amplifier for Physiological Instrumentation," J. R. Smith, Jr., Ames Res. Ctr.

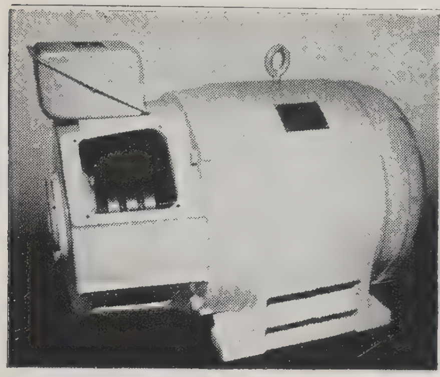
(Continued on page 78A)

Vol. 2, No. 4

Nickelonic News

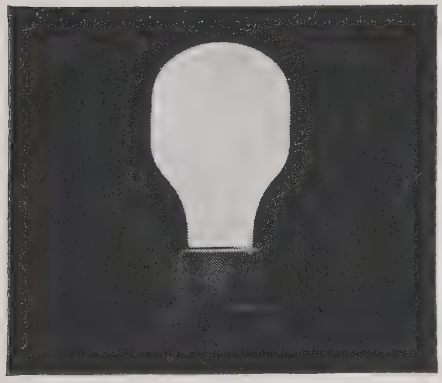
DEVELOPMENTS IN NICKEL AND NICKEL ALLOYS AND THEIR APPLICATIONS

5 MATERIALS PROBLEMS SOLVED BY 5 HUNTINGTON ALLOYS



Keep it low, keep it uniform

Low wear, uniform wear are important requirements in these slip rings—constantly subject to electrical and mechanical erosion by the brushes in this new 60 HP 1800 rpm wound rotor motor. Monel* nickel-copper alloy was chosen for slip rings in this application by General Dynamics' Electro Dynamic Division, Bayonne, N. J. because of its hardness, uniform throughout the ring.



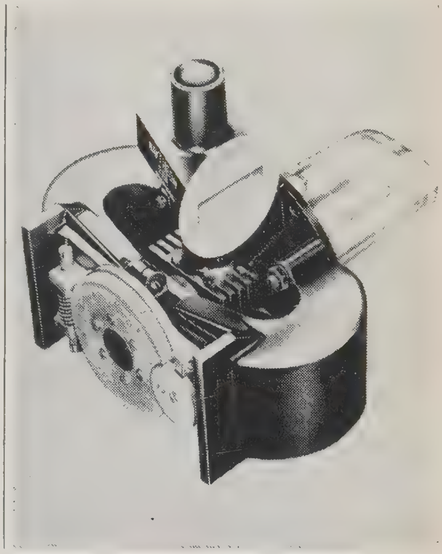
Arc triumph

If the filament fails in this argon-filled Eye Saving lamp, the resulting arc might well burn through the base. To prevent this, engineers at Westinghouse Lamp Division, Bloomfield, N. J., provide a fuse of "D"* Nickel. Its superior oxidation resistance, hardness and strength make possible an efficient fuse of wire only .008 to .010 inches in diameter.

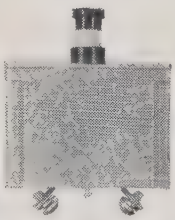
A vile base problem solved

Engineers at Raytheon Company, Waltham, Massachusetts, solved a design problem by making the cathode base of their RK2J51 magnetron from "A"* Nickel Electronic Grade. The favorable electrical and mechanical properties of this alloy made it their choice for the cathode of this magnetron . . . an important component in many airborne radar systems.

"A" Nickel Electronic Grade is appropriate for a wide variety of electronic and electrical applications. Its resistance to corrosion is of a high order. Oxidation at high temperatures is low and any oxide scale formed is tightly adherent.



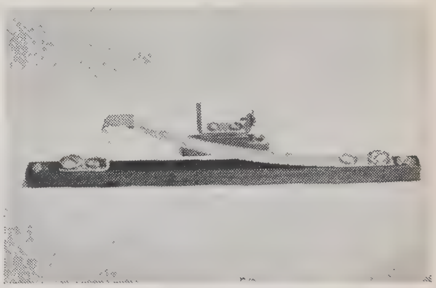
Don't try to magnetize Monel "403"



Regardless of form, size or temper, its permeability won't exceed 1.1 at 27°F in a field strength of 0.5 oersted. That's why Wood Electric Company, Lynn, Massachusetts, uses Monel* "403" nickel-copper alloy for terminals in this thermal circuit breaker for aircraft. This useful alloy also resists atmospheric corrosion, stays bright and clean in service.

Huntington Alloys

The helpful booklet "Inco Nickel Alloys for Electronic Uses" describes many suitable Huntington Alloys. For a copy, and for help with materials problems, write today to:



Burglar's bane in sun or rain

Used in an alarm system made by American District Telegraph Company, New York, this vital part must resist weather, corrosive atmospheres, and wear. For necessary physical and electrical properties—including excellent corrosion resistance—spring is a strip of annealed, round edge spring temper Inconel* nickel-chromium alloy; contact post is cold rolled Monel nickel-copper alloy strip.

*Inco trademark

INCO TRADE MARK

HUNTINGTON ALLOY PRODUCTS DIVISION
The International Nickel Company, Inc.
Huntington 17, West Virginia

HUNTINGTON ALLOYS

Professional Group Meetings

(Continued from page 76A)

San Francisco—January 18

"Hospital Automation—Progress and Prospects," Dr. M. Blumberg, Stanford Research Institute, Menlo Park, Calif.

Washington, D. C.—February 23

"The Electrical Basis of Neurophysiology," Dr. K. Frank, National Institute of Neurological Diseases and Blindness, Bethesda, Md.

Broadcast and Television Receivers

Chicago—October 14

"Transistorized Stereo Amplifier," J. Dols, Warwick Mfg. Corp., Chicago, Ill.

Circuit Theory

Omaha-Lincoln—March 2

2nd Annual IRECTIONS (IRE Circuit Theory Iowa-Nebraska Symposium).

"Similarities in Wave Behavior" (Bell Laboratories film narrator J. Shive).

"Wave Propagation in the Telephone Industry," R. E. Starnes, Northwestern Bell.

"Wave Propagation in the University of Nebraska Curriculum," Dr. W. C. Robison, Univ. of Neb.

Presentation: Science Teacher Award to H. Dally, Omaha North High School.

Communications Systems

Chicago—January 13

"A Random Axis Communication System," R. W. Dronsuth, Motorola, Inc., Chicago, Ill.

Los Angeles—October 25

"The Role of the Jet Propulsion Laboratory for Project Echo," W. K. Victor, J.P.L., Pasadena, Calif.

Toronto—February 27

"Canadian Topside Sounder Satellite," Dr. R. C. Langille, Defense Research Board, Ottawa.

Washington, D. C.—February 13

"Digital Television," W. C. Coombs, National Bureau of Standards, Boulder, Colorado; Dr. R. C. Webb, Colorado Res. Corp.; and R. G. Salaman, Ball Bros. Res. Corp., Boulder, Colo.

Communications Systems Information Theory

Los Angeles—January 24

"A Survey of Space Communications," J. E. Taber, Space Technology Labs., Los Angeles.

"Chapter 1 of Shannon Revisited," S. W. Golomb, Jet Propulsion Lab., Pasadena.

(Continued on page 81A)

Specify the new Amperex® P·A·D·T 40

the 2 times faster PNP Germanium Switching Transistor

$$* \; U = \frac{1}{\text{nanoseconds x pennies}}$$

Right from the sketch-pad stage, plan your computer switching circuits with the new PADT-40.* The extreme speed and efficient design of the PADT-40 gives more U (usefulness factor) and lower cost x switching time. This results in fewer transistors to buy, less complicated circuits to design, and the elimination of many costly components because of multi-function circuit usage. But speed, of course, is only one of the cost-and-production advantages inherent in the PADT-40; **RELIABILITY**, as only the revolutionary Post Alloy Diffusion Technique can provide, is another; **AVAILABILITY**, as only the mass-production techniques employed at the new Amperex plant in Slatersville, R. I., can provide, is still another; **LOW PRICES** (no higher than for low-speed transistors) . . . plus **INTERCHANGEABILITY** with many conventional mesa transistors, round out our 'package'. Yes, the new Amperex PADT-40 is truly worth specifying . . . *now!*

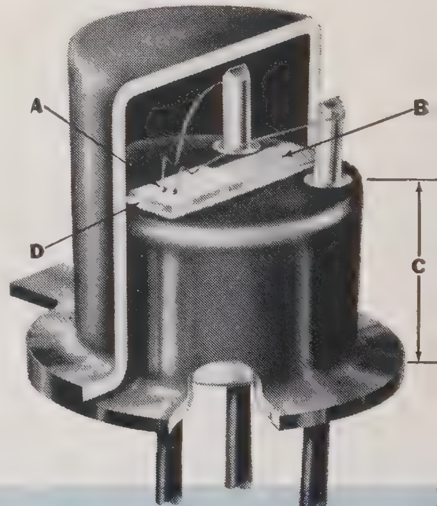
High Speed, plus . . .

MECHANICAL RUGGEDNESS — guaranteed by the *only* process that combines the best qualities of both the alloy and the diffusion methods. As a result, the PADT-40 is resistant to vibration and shock.

PADT RELIABILITY — Hermetically sealed in a standard TO-18 case, the PADT-40 has a deep diffused and extremely thin active base region. As a result, the h_{FE} and switching time are virtually independent of surface effects and temperature changes.

- **A** *Rugged, mechanically reliable eutectic solder joints*
- **B** *Flat bed attachment for good heat dissipation*
- **C** *Long path prevents weld contamination of transistor*
- **D** *Gold doped for high speed*

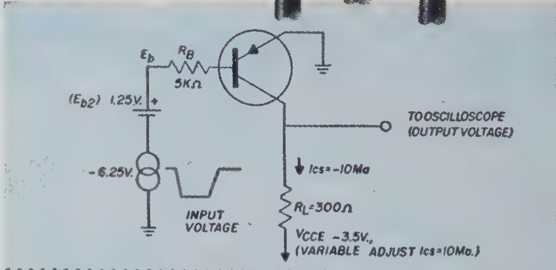
- *Extremely high cut-off frequency*
- *High Beta* • *Low resistivity germanium*



TOTAL SWITCHING TIME

including Rise, Fall, Delay and Storage

...130 NANOSECONDS!



AVERAGE SWITCHING TIMES

Value	t_d	t_r	t_s	t_f	t_{tot}	
Time	30	20	40	40	130	Nanosecond

ask Amperex *for complete data and new transistor brochure*

AMPEREX ELECTRONIC CORPORATION
230 Duffy Avenue, Hicksville, L. I., N. Y.

In Canada: Rogers Electronic Tubes & Components, 116 Vanderhoof Ave., Toronto 17, Ont.

Professional Group Meetings

(Continued from page 78A)

Component Parts

Los Angeles—February 13

"Modern Engineering and Manufacturing Techniques for Reliability in Relays," W. Sellars, Leach Corp.

Philadelphia—January 24

"High Density Packaging," E. A. Szukalski, RCA.

Electron Devices

Boston—February 2

"Epitaxial Diffused Transistors," Dr. H. H. Loar, Bell Telephone Labs., Murray Hill, N. J.

New York, Long Island, and Northern New Jersey—January 19

Dr. R. Adler, Zenith Radio Corp., Chicago, Ill., was the featured speaker.

Electron Devices/Microwave Theory and Techniques

San Francisco—February 15

"Report on Stanford's Project M—The Two-Mile Linear Electron Accelerator," Dr. K. Mallory, Stanford University.

Electronic Computers

Akron—January 17

"Achieving a Reliable System," G. Raymond, Remington Rand, Minneapolis.

Binghamton—March 1

"Cryogenics—In General," A. Juster, Servomechanisms, Inc.

Detroit—January 30

"Use of Automatic Data Reduction Techniques in Study of Physiological Aspects of Emotion," Dr. A. F. Ax and G. Zacharopoulos, Lafayette Clinic, Detroit, Mich.

Fort Worth—January 23

"RH-2 Navigation and Flight Computer," R. D. Watson, Bell Helicopter Corp., Hurst, Tex.

Los Angeles—February 16

"UCLA's Fixed Plus Variable Computer," Dr. G. Estrin, R. Turn and M. Aoki, UCLA Engineering Dept.

New York and Northern New Jersey—February 2

"Magnetic Circuits for Logic and Storage," Dr. E. E. Newhall, Bell Telephone Labs., Murray Hill.

Omaha-Lincoln—February 28

"IRE Proceedings Computer Issues," October, 1953 and January, 1961, C. Walston, IBM, Omaha.

"What the Young Engineer Should Know About Reliability," M. A. Young, IBM, Owego, N. Y.

"Display of Older and Recent Computer Plug-In Units," J. R. Wignall, IBM, Lincoln.

Philadelphia—February 7

"The Design and Manufacture of the High Speed Switching MADT Transistor," C. D. Simmons, Philco Corp., Lansdale, Pa.

Pittsburgh—February 23

"Present and Future Trends in Computers," M. Rubinoff, Univ. of Pa.

San Francisco—January 24

"Finite State Machines—The State of the Art," A. Gill, University of California.

Washington, D. C.—March 1

"Fosdic III, an Electro-Optical Scanner for Processing the 1960 Census Data," M. L. Greenough, National Bureau of Standards, Washington, D. C., and E. S. Stein, Rabinow Engineering Co., Washington, D. C.

Washington, D. C.—January 31

"Four Layer Devices and Their Application to Digital Techniques," E. Jackson, Transitron Electronic Corp., Wakefield, Mass.

Engineering Management

New York—February 23

"Application of PERT to Polaris Program," K. Williams and D. Goff, Sperry Gyroscope, Great Neck, L. I.

Rome-Utica—February 21

"Our Growing Stature in the Space Age," J. Bridges, Dept. of Defense, Wash., D. C.

Rome-Utica—January 17

"Saving Dollars by Non-Linear Modeling," and "My Trip to Russia," Prof. K. M. Siegel, University of Michigan, Ann Arbor.

Rome-Utica—September 13

"The Engineering Manager, a Dependent Variable," G. Frantz, MITRE Corp.

Syracuse—January 31

"Information Retrieval," C. D. Gull, GE, Washington, D. C., and Dr. F. R. Whaley, Tonawanda Labs., Linde Co., Tonawanda, N. Y.

Engineering Management Reliability and Quality Control

San Francisco—January 23

"Components for Reliable Design," J. Lavrischeff, Lawrence Radiation Lab., Berkeley, Calif.

(Continued on page 82A)

Professional Group Meetings

(Continued from page 81A)

Engineering Writing and Speech

Boston—February 21

"Optical Scanning—Its Uses in Information Retrieval, Transfer and Indexing," G. L. Fischer, Farrington Electronics Inc., Needham, Mass.

New York—June 16

"Workshop Session," Profs. Barr, Miller and Holub of Columbia and Fairleigh Dickenson Universities.

"Is Reading Necessary?" E. E. Grozdar, Electronic Design, New York, N. Y.

New York—June 14

"Choosing Proper Words When Writing and Speaking," R. W. Holub, Fairleigh Dickenson Univ., Teaneck, N. J.

New York—June 9

"Speech and Mannerisms," M. Miller, Fairleigh Dickenson University, Teaneck, N. J.

New York—June 7

"Definition of the Written and Spoken Word," D. Barr, Columbia University, New York, N. Y.

Information Theory

Boston—February 9

"Seismological Data Collection and Processing," Dr. B. P. Bogert, Bell Telephone Labs., Murray Hill, N. J.

Instrumentation

San Francisco—February 7

"State of the Art in Precise Phase Measurement from Very Low To High Frequencies," Dr. P. Yu, Ad-Yu Electronics Lab., Passaic, N. J.

Washington, D. C.—February 6

"Fluid Amplification through Stream Interactions," B. M. Horton and R. Warren, Diamond Ordnance Fuze Labs., Washington, D. C.

Microwave Theory and Techniques

New York—November 17

"The Modern Radar System," Dr. J. S. Burgess, Rome Air Development Center.

New York—October 6

"Electronically Steered Antennas," H. Shnitkin, W. L. Maxson Corp., New York, N. Y.

Schenectady—January 25

"Resistance Noise Calculation," W. Mumford, Bell Telephone Labs., Whippany, N. J.

Military Electronics

Boston—December 7

"Space Surveillance Techniques Used at Space Surveillance Control Center," E. W. Wahl, NASCC, Hanscom Field, Bedford, Mass.

Washington, D. C.—January 10

"Recent Results of Upper Atmosphere Research," Dr. H. Friedman, U. S. Naval Res. Lab.

Washington, D. C.—November 28

"Motivation for Research," Dr. R. M. Page, USNRL, Washington, D. C.

Military Electronics Antennas and Propagation

Washington, D. C.—February 21

"The Determination of the Thermodynamic and Electrodynamic Constants of the Surface of the Moon," K. M. Siegel, University of Michigan, Ann Arbor.

Military Electronics Engineering Management Engineering Writing and Speech

Washington, D. C.—January 30

"Psychology of Selling Ideas to Management," T. Gays, Leadership Training Institute of D. C., Washington, D. C.

Nuclear Science

Los Angeles—February 21

"Hot Cell Operation," J. Leivch, Atomics International, Canoga Park, Calif.

Oak Ridge—February 16

"Japan, Its People and Its Industry," E. Arakara, ORNL.

Product Engineering and Production

Los Angeles—January 25

"Microelectronics," L. Katzin and Leopold Cann, Ramo-Wooldridge, Canoga Park.

Los Angeles—January 9

"Thin Films," Dr. L. J. Varnein, Jr., Bell Telephone Labs.

"Thin Films," Ben Solow, International Resistance Co.

New York—December 7

"Value Engineering in Practice," M. Kaplan, Loral Electronics Corp., Bronx, N. Y.

"Yard Stick Techniques for Optimum Weight, Volume and Mechanical Reliability," D. Ehrenpreis, Independent Consultant, New York, N. Y.

(Continued on page 86A)

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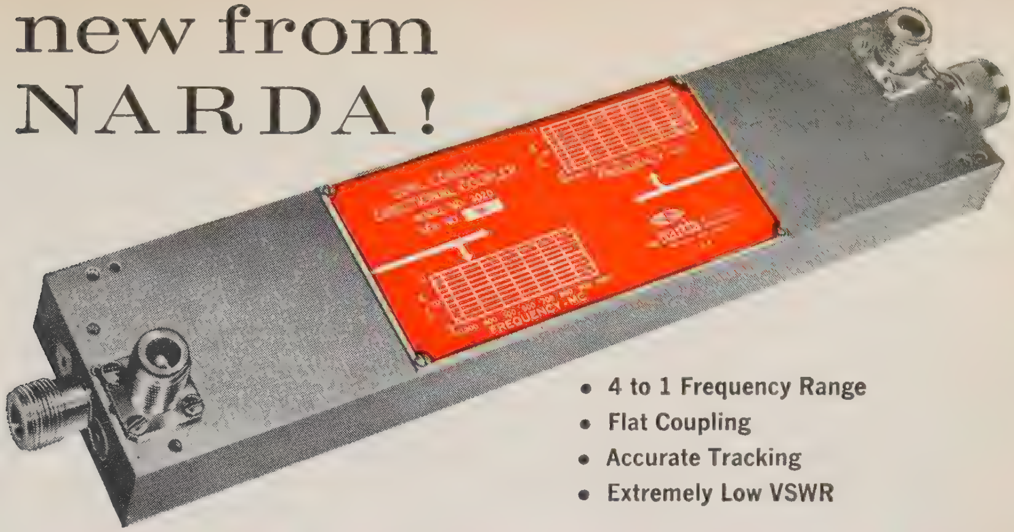

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Max VSWR—main line	1.05	1.10
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Price	$200.	$185.

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Professional Group Meetings

(Continued from page 83A)

Reliability and Quality Control

Los Angeles—February 20

"Reliability in Design," B. H. Hershkowitz, North American Aviation, Downey, Calif.

"Integrated Reliability Program," R. H. Hover, North American Aviation.

"Combined Environmental Tests for Reliability Assessment," D. L. Rowlands, North American Aviation.

Philadelphia—February 7

"Maintainability Measurement and Evaluation," R. Redfern, IBM, and J. McKendry, HRB Singer, Inc.

Debate between above speakers, and question and answer session.

San Francisco—January 10

Installation of officers, discussion of by-laws and operating procedures, and planning activities for coming year.

San Francisco—January 23

"Reliability from the Design Engineer's Viewpoint," J. Lavrischeff, Lawrence Radiation Labs., Berkeley, Calif.

San Francisco—December 21

"Reliability through Monitoring and Redundancy," R. McDonald, Sylvania Electric Products Inc., Mountain View, Calif.

Space Electronics and Telemetry

Albuquerque—February 21

"Logic Modules for Signal Conditioning on Space Radiation Experiments," F. E. Thompson, Sandia Corp., Albuquerque, N. M.

Vehicular Communications

Detroit—February 22

"Optimizing Watts, Decibels, Dollars," A. C. Giesselman, General Electric Co., Lynchburg, Va.

Los Angeles—February 16

"Space Systems—Past, Present and Future," R. W. Hallet, Jr., Douglas Aircraft Co., Santa Monica, Calif.

Membership

The following transfers and admissions have been approved and are now effective:

Transfer to Senior Member

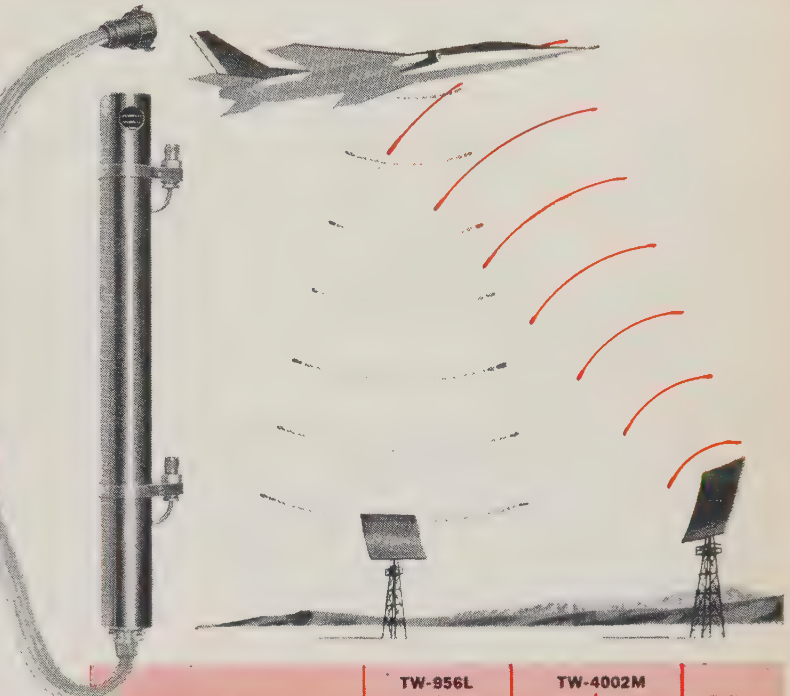
Aagaard, J. S., Evanston, Ill.
Barr, W. W., Sr., Dallas, Tex.
Birx, D. L., Philadelphia, Pa.
Blake, L. V., Washington, D. C.
Bobyn, E. J., Dean Haag, Netherlands
Curtis, G. D., Irving, Tex.
Davenport, E., Auckland, New Zealand
Davis, R. J., Dewitt, N. Y.
Dowe, R. J., New Orleans, La.
Grushkin, M., Sunnyvale, Calif.
Hall, R. E., Anaheim, Calif.
Henley, E. J., Fanwood, N. J.
Hensperger, E. S., Metuchen, N. J.
Hoff, H B., Fairview Park, Ohio
Hurley, L., Shreveport, La.
Karmiol, E. D., Rosemont, Pa.
Kellway, B. J., Belfast, North Ireland
Larsh, A. E., Jr., Richmond, Calif.
Lazinski, R. H., Glenside, Pa.
Lomaz, D. d., Cleveland, Ohio
Madsen, E. C., La Mesa, Caifi.
Marsocci, V. A., Flushing, L. I., N. Y.
Miyata, J. J., Montery Park, Calif
Morris V. B., Jr., Baltimore, Md.
Moseley, J. A., Santa Barbara, Calif.
O'Bryan, R. L., Oxnard, Calif.
Perry, F. M., Utica, N. Y.
Ratz, H. C., Saskatoon, Sask., Canada
Schrader, G. F., Lakewood, Calif.
Taub, J. J., Plainview, L. I., N. Y
Vall, A E., Rolling Hills Estates, Calif.
Van Nice, R. I., Glenshaw, Pa.
Whitchurch, N. E., Akron, Ohio
Whiting, W. E., Bakersfield, Calif.

Admission to Senior Member

Barnes, R. F., Scottsdale, Ariz.
Barto, J. P., Chicago, Ill.
Benton, B. M., Bellevue, Wash.
Best, J. H., Towson, Md.
Brandell, S. R., Brooklyn, N. Y.
Brown, A. M., Ventura, Calif.
Brown, K., Los Angeles, Calif.
Burns, C. H., Cleveland, Ohio

(Continued on page 88A)

MICROWAVE DEVICE NEWS from SYLVANIA



NEW! GRIDDED TWT's for PULSED or CW operation in S band frequencies

CHARACTERISTICS	TW-956L Min.	TW-956L Max.	TW-4002M Min.	TW-4002M Max.	Units
Cathode Current	—	50	—	4.5	mAdc
Helix Current	—	7	—	3.2	mAdc
Grid Current	—	7	—	1.0	mAdc
CW RF Power Output (sat)	2W	—	10mW	—	
Small Signal Gain					
(−30dbm input)	37	—			db
(−40dbm input)			35	—	db

Designed for Airborne ECM . . .

Sylvania TW-956L, TW-4002M are PPM-focused, magnetically shielded, weigh only 4 lbs., are just 15″ long, 1.4″ in diameter, temperature compensated for −65° to +72° C.

Utilizing a unique TWT design incorporating grids, both types exhibit sharp cutoff characteristics. They feature relatively flat frequency response over the full 2.0–4.0 kMc frequency range.

TW-956L is capable of 2W CW saturated power output. TW-4002M features CW saturated power output of 10mW. Both types can be provided with virtually any mounting.

For full data on these types contact your Sylvania Sales Engineer, or write Electronic Tubes Division, Sylvania Electric Products Inc., Dept. MDO-E, 1100 Main St., Buffalo 9, New York.

FIRST DC BLOCK COVERING 2.5 TO 10,000 MC

Sylvania SB-100, wideband coaxial device providing dc and low frequency isolation, features low VSWR of 1.3 : 1 or less over its rated frequency range and maximum insertion loss of 0.2db. CW power rating is 100W, peak power is 2KW, voltage isolation of 2KV dc. SB-100 offers excellent stability over a temperature range of −50° to +100° C. It is available with type N connectors. Soon to be available are the SB-101 having BNC connectors and the SB-102 featuring TNC connectors.

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GENERAL TELEPHONE & ELECTRONICS

GENERAL SYSTEM

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(Continued from page 86A)

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(Continued on page 90A)

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(Continued from page 88A)

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(Continued on page 92A)

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(Continued from page 90A)

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O'Leary, W. J., Merchantville, N. J.
O'Neil, J. M., Rome, N. Y.
O'Niel, T. J., Kensington, Md.

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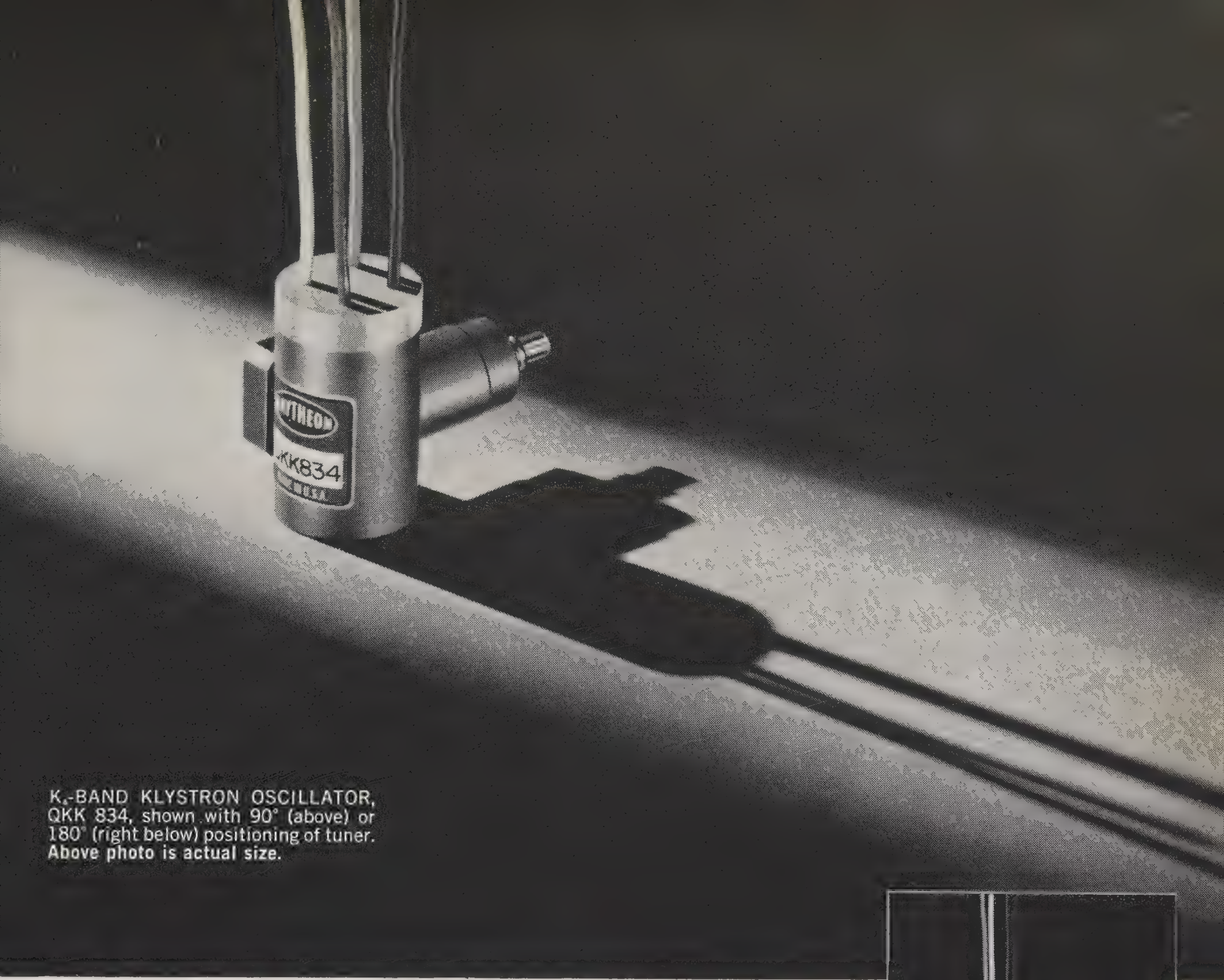
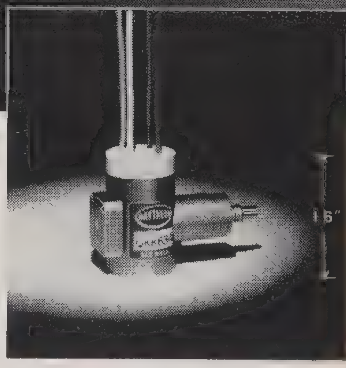


K_a-BAND KLYSTRON OSCILLATOR, QKK 834, shown with 90° (above) or 180° (right below) positioning of tuner. Above photo is actual size.



New klystrons hold characteristics in grueling aerospace environments

K_a- and K-band tubes are tunable from 34.0 – 35.6 and 23.5 – 24.5 kMc

Now, Raytheon combines the advantages of small size, extreme ruggedness, thermal stability, and smooth wide-range tunability in a 20mW reflex klystron.

The new QKK 834 for K_a band and QKK 923 for K band are all ceramic and metal tubes with typical electronic tuning range of 110 Mc. The tuner, utilizing a sapphire rod, can be specified for positioning anywhere on the circumference of the resonator at least 90 degrees from output flange (see illustrations above). Write today for detailed technical data or application service to Microwave & Power Tube Division, Raytheon Company, Waltham 54, Massachusetts. In Canada: Waterloo, Ontario.

QKK 834, QKK 923 — GENERAL CHARACTERISTICS	
Power Output	20 mW (nominal)
Frequency	34-35.6*; 23.5-24.5† kMc
Resonator Voltage	400 V
Reflector Voltage Range	—65 to —175V
Temperature Coefficient	± 0.5 Mc/°C
Cooling	convection (no blower needed)
Overall Dimensions	1 5/8 x 1 1/16 x 2 in.*

*QKK 834 †QKK 923

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38 Transactions, *6, & *9, and Vol. 2, No. 1-3; Vol. 4, No. 1, 2, 3; Vol. 5, No. 2, 3, 4; Vol. 6, No. 1, 2, 4; Vol. 7, No. 1, 2, 3. 4.

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May, 1961
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Proceedings of the IRE

Poles and Zeros

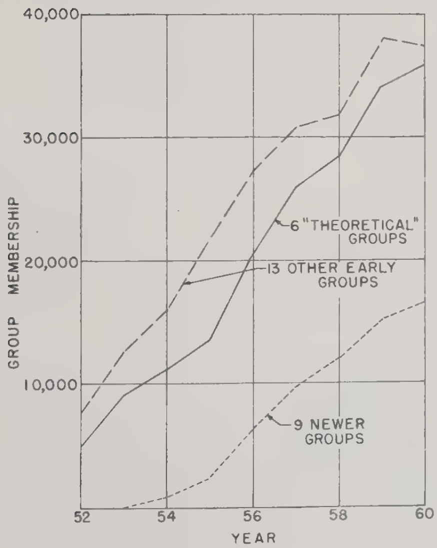
Professional Groups. The following information on the IRE Professional Groups has been prepared for Poles and Zeros by Ernst Weber, Past President of IRE, presently serving as Chairman of the Professional Groups Committee.

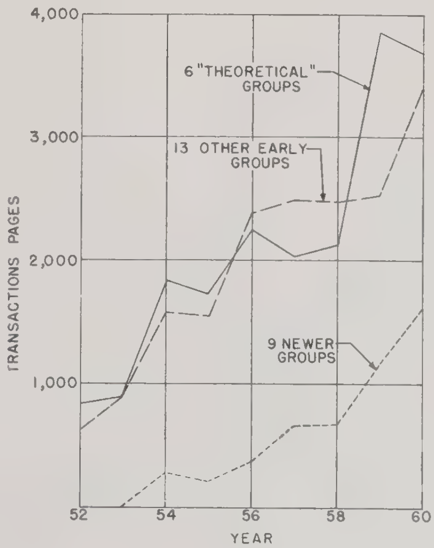
"One of the major factors in the phenomenal growth of IRE to the status of the largest engineering society in the world is, undoubtedly, the system of professional groups. Started in 1948 it provided "unity through diversity," thus preserving the single society concept which gave strength and maturity yet permitting groupings according to professional interests with autonomy of planning and publishing. The system has been a vital force in bringing to the members the full impact of the dynamic progress of science and engineering and in stimulating their own professional growth.

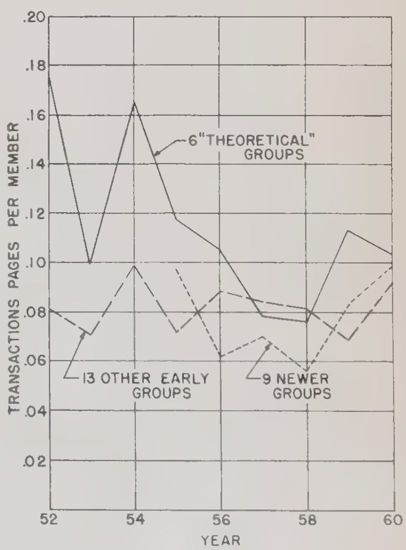
"The strongest membership services that the professional groups perform are technical conferences or symposia, and the publication of TRANSACTIONS. The basic autonomy of the groups permits publication of TRANSACTIONS issues at various periodic intervals and either in letter press or offset printing. The external format has been standardized to facilitate library handling, but each group has distinctive color combinations on the cover pages (an attractive spectral array) once more illustrating the motto: "unity through diversity."

"The graphs below show that the total number of pages in the TRANSACTIONS has generally kept pace with the growth of the group membership; the groups publish nearly one page per 10 members. The corresponding financing problem is as severe as for any professional publication of high quality. To assist the professional groups without interfering with their autonomy, the IRE Board of Directors originally allocated funds from the over-all budget in relation to group membership. Because of the importance of the TRANSACTIONS, the fund allocation was changed on January 1, 1959, to be equal to one-third the publication budget of the individual group as agreed to by the Executive Committee. The beneficial effect of this changed policy is impressively illustrated in the graph.

"An interesting and quite significant aspect in the development of the professional group system has been the fact that, for the last ten years, six of the total of 28 groups have constituted 40 per cent of the total membership and have contributed from 40 per cent to 50 per cent of all TRANSACTIONS pages, and this in spite of the fact that the actual number of groups increased from 19 to 28 in that same period. The even more astonishing fact is that these six groups can be characterized as the "theoretical groups," concentrating their attention on the scientific principles underlying the newest developments of fundamental nature in which good publications are of the essence. In the order of membership, these groups are: Electronic Computers, Circuit Theory, Electron Devices, Microwave Theory and Techniques, Antennas and Propagation, and Information Theory. It has always been characteristic of IRE to integrate science and engineering; the strength of the science oriented groups attests that IRE thus performs a distinctive membership service."—F. H., Jr.







E. Finley Carter

Director, 1961–1963

E. Finley Carter (A'23–F'36) is President of the Stanford Research Institute, Menlo Park, Calif., and a member of its Board of Directors. He has had a distinguished career in research and management, and was a pioneer in the field of human relations.

A native of Elgin, Tex., he received the B.S. degree in electrical engineering (With Distinction) from Rice Institute, Houston, Tex., in 1922.

He began his career as a student engineer with the General Electric Company's Radio Division in Schenectady, N. Y. He was soon advanced to the position of section leader, where he helped engineer the design of several major radio broadcasting stations. He later became a division engineer of the Radio Department. When he joined the United Research Corporation, New York, N. Y., in 1929, he was placed in charge of the Radio Division.

After three years in New York, he joined Hygrade Sylvania Corporation (later Sylvania Electric Products, Inc.) in Emporium, Pa., as a division engineer. His long interest in the field of human relations resulted in his appointment as Director of Industrial Relations of Sylvania in 1941, with headquarters in New York, N. Y. This entailed a move to Manhasset, N. Y., where he lived for the next thirteen years. During this time he helped formulate and implement the corporate philosophy of decentralization and multiple-plant operations pioneered by Sylvania. He initiated progressive programs in industrial relations and organization development. In 1945 he became Vice President in charge of Industrial Relations. A year later he was appointed Vice President in charge of Engineering, and in 1952 he became Vice President and Technical Director.

He joined the Stanford Research Institute in 1954 as the Manager of Research Operations. In 1956 he became Director of the Institute, as well as a member of SRI's Board of Directors. In March, 1959, he was named as SRI's first president.

He has been granted over 20 patents in radio and carrier current communication, remote control, telemetering, automatic synchronization, time-delay relays and vacuum-tube devices. Among his major inventions were two dealing with single frequency duplex systems, and one on automatic volume control. A number of his basic discoveries are in common use today in radio communication and control systems.

Mr. Carter is an active ham radio operator (K6GT), and is a member of several professional associations, including the American Institute of Electrical Engineers and the Illuminating Engineering Society. He is also a member of the honorary engineering fraternity, Tau Beta Pi, the Scientific Research Society of America and the American Radio Relay League. He is a Registered Professional Engineer in the state of New York.

Scanning the Issue

Low-Noise Amplifiers for Centimeter and Shorter Wavelengths (Wade, p. 880)—Reference is often made to the fact that radio and electronics have greatly extended man's five senses during the past half century. As one example, just in the last decade the development of nearly noiseless devices has increased the responsiveness of his electronic sensory receptors by an additional three orders of magnitude. Ten years ago the only low-noise microwave amplifiers were a few traveling-wave tubes having noise temperatures of about 3000°K. Today we find that these noise temperatures have been reduced to as low as 250°K. Even more important, we now have at hand an impressive array of new low-noise devices, including electron-beam and solid-state parametric amplifiers, tunnel diodes, masers, irasers and lasers. Noise temperatures of only 25°K are now attainable with amplifiers operating at room temperature, and 10°K with refrigerated devices. As for the future, attention is beginning to turn to photon counters and photosensitive detectors, which are inherently even quieter than masers. Every member will want to read this excellent description and state-of-the-art survey of recently developed low-noise techniques that are of such dominant importance today.

General Relativity for the Experimentalist (Forward, p. 892)—The advent of the space age has brought with it a renewed interest in relativity theory. It has now become possible to place apparatus in extraterrestrial space, thus opening important new avenues for testing the effects of the earth's gravitational field. Not only do we have new vantage points from which to conduct relativity experiments, but new and better tools, such as atomic clocks, by which to perform them. The task of finding experiments that are suitable, however, is a formidable and discouraging one. Not only are the mathematics of the problem very specialized and difficult, but the laborious calculations almost always lead to the discovery that the relativistic effect under study is too small to be observed. In this paper the author develops a simplification of Einstein's theory of general relativity which permits him to recast the problem in more familiar terms by means of an analogy between electromagnetism and gravitation. Not only does this provide a mathematical shortcut but it also makes it possible for anyone versed in electromagnetic theory to determine with reasonable accuracy whether the effect under study is likely to be large enough to be of experimental interest and, hence, warrant more rigorous study. For the general reader, the electromagnetic analogies and the examples given provide an interesting insight into the consequences of general relativity. For the electromagnetic specialists in the audience there is an interesting offer from the author (see footnote 3 of the paper) plus the possibility that one of you with a practical turn of mind will think of an experiment to detect the gravitational equivalent of the magnetic field.

A Matched Amplifier Using Two Cascaded Esaki Diodes (Hamann, p. 904)—A new type of circuit has been developed for obtaining amplification with negative resistance devices which employs two negative resistance elements instead of one. This arrangement makes it possible to match the amplifier to both the source and load. It also makes it easier to achieve a low noise figure. Although the circuit described here employs tunnel diodes, the technique appears to be equally promising for parametric amplifiers and other negative resistance devices.

Network Synthesis with Negative Resistors (Carlin and Youla, p. 907)—Negative resistance, which figures prominently also in two of the preceding papers in this issue, can no longer be regarded as a special and unusual phenomenon. The recent development of devices such as variable-capacitance diodes and tunnel diodes has made the negative resistance element a common and important part of modern electronic circuitry. It is not surprising, therefore, to find in this paper that circuit theorists, too, have now added the negative resistor to the five basic circuit elements which they traditionally employ in problems of linear network analysis and synthesis. The adoption of this new building block is an event of more than ordinary significance. The consequences of its entry into the network theory domain, as revealed in this paper, will have broad implications for the whole field of linear active networks.

Fluctuation Noise in Semiconductor Space-Charge Regions (Giacoletto, p. 921)—Reference to Fig. 6 of the first paper in this issue will show that parametric diodes, when refrigerated, can be operated with noise temperatures nearly as low as those of masers. The interesting and important question arises: what is the minimum level of noise inherent in a parametric semiconductor diode? As yet, it has never been measured because of external circuit noise. This paper predicts that at low enough temperatures the external circuit noise can be reduced to a point where internal noise in the junction dominates and becomes the ultimate limiting factor, and that with sufficient care it can be measured.

Electrodeless Measurement of Semiconductor Resistivity at Microwave Frequencies (Jacobs, *et al.*, p. 928)—A new technique for measuring the bulk resistivity of semiconductors has been developed which avoids the problems associated with ohmic contacts of electrodes. The method, which requires no electrodes, makes use of the absorption of microwave power by the material to arrive at a direct measurement, tying in nicely with a similar method for measuring lifetime reported here last year.

Accuracy and Limitations of the Resistor Network Used for Solving Laplace's and Poisson's Equations (Hechtel and Seeger, p. 933)—This paper brings to the fore an attractive analog method, first proposed 18 years ago, for solving two types of equations which are widely encountered by engineers and physicists today. It will be of particular interest to those engaged in electron-optical problems, since the method shows interesting advantages over the more common electrolytic-tank analogs now in use.

Optimum Capacitor Charging Efficiency for Space Systems (Mostov, *et al.*, p. 941)—This paper is concerned with methods of periodically charging a capacitor bank, as is required in many pulsed-operation devices such as plasma engines, without the usually accepted 50 per cent loss of energy to ohmic dissipation. It is shown that the efficiency of energy transfer can be substantially improved without resorting to inductors that are too heavy by employing proper voltage shaping techniques.

A Gas-Discharge Microwave Power Coupler (Olthuis, p. 949)—The author has devised a cavity that supports two resonant modes at the same frequency so that power may be exchanged between them in a controlled fashion by means of a controllable gas discharge and controllable magnetic fields within the cavity. Because the mechanism can be externally controlled, it offers a number of interesting possibilities as a microwave device, including a power switch, a power divider and a modulator.

1960 Index to Abstracts and References—The entries in Abstracts and References which appeared at the rear of the editorial pages each month in 1960, in addition to listing many IRE articles, covered over 4000 other contributions in technical publications all over the world. An index to this wealth of material appears as Part 2 of this issue.

Scanning the Transactions appears on page 983.

Low-Noise Amplifiers for Centimeter and Shorter Wavelengths*

G. WADE†, SENIOR MEMBER, IRE

Summary—**A number of new techniques for detecting signals at centimeter and shorter wavelengths are making possible far better noise performance than ever before attainable. Over the past ten years, noise temperatures have been reduced from around 3,000°K to less than 5°K. The low-noise devices which are currently competitive or which show future promise include traveling-wave tubes, parametric amplifiers, tunnel diodes, masers, photon coun ers, and photosensitive detectors. This paper discusses the various techniques for attaining low noise and summarizes the achievements relative to each technique.**

Effective noise reduction in traveling-wave tubes is accomplished by making the beam flow through an extended low-velocity region. Noise temperatures as low as 250°K have been measured on traveling-wave tubes at *S* band. Low noise has been attained in parametric devices using both electron beams and semiconductor diodes. Some of the lowest noise temperatures measured on any unrefrigerated microwave amplifiers are those for electron-beam parametric amplifiers. Refrigerated semiconductor-diode parametric amplifiers have given even lower noise temperatures. The tunnel-diode amplifier is the most recent negative resistance microwave device. It seems to have limitations as far as extremely low-noise performance is concerned, but nevertheless it is a good competitor for many low-noise applications. So far, the maser is the best in microwave low-noise amplification. Intrinsic maser noise temperatures of a few degrees are readily attainable. Photon counters and photosensitive detectors are inherently even quieter and give promise of considerable future potential.

Introduction

Over the past several years, a number of new techniques have been discovered for low-noise detection of microwave signals and higher frequency signals. The new techniques are making possible far better sensitivities than previously attainable. The discovery of these techniques has been timely, because simultaneously with their discovery has developed an urgent need for better sensitivity. A few of the many applications which demand low-noise detection are:

1) Early-warning radars and satellite-based radars,
2) Passive detection and mapping,
3) Intercontinental communication by means of active or passive satellite repeaters,
4) Radio astronomy.

Ten years ago the only low-noise microwave amplifiers were a few traveling-wave tubes built in various laboratories. Their noise temperatures were around 3000°K.[1,2] At the present time there are a number of new devices capable of substantially better performance. Some of them are now out of the laboratory stage and are commercially available. Noise temperatures have been reduced by two and three orders of magnitude. A list of low-noise devices which are currently competitive or which show future promise is given below:

1) Traveling-wave tubes (using new techniques for beam noise reduction),
2) Parametric amplifiers,
3) Tunnel diodes,
4) Masers,
5) Photon counters and photosensitive detectors.

This paper will discuss the various techniques for achieving low-noise detection and summarize the state of the art relative to each technique.

Low-Noise Traveling-Wave Tubes

The traveling-wave tube is the old stand-by in low-noise microwave amplification. Important new techniques have been discovered recently for reducing its beam noise. The noise temperatures now obtainable are an order of magnitude better than those of ten years ago.

In its simplest form, the traveling-wave tube consists of a pencil-shaped electron beam traveling near a slow-wave structure, usually a helix. The electrons of the beam interact continuously with a traveling electromagnetic wave on the slow-wave structure to produce the amplification. The interaction extends over the entire length of the structure and therefore lasts for a relatively long period of time. This fact makes possible high amplification with a nonresonant structure of relatively low impedance. As a consequence of the nonresonant structure, it is possible to obtain extremely wide bandwidth in a traveling-wave tube. The characteristics of wide bandwidth and low noise make the traveling-wave tube particularly attractive for passive detection and radio astronomy.

Let us consider briefly the amplification process in order to understand the noise characteristics. The wave picture tells a good deal about the amplification process.[3] The electron beam as well as the slow-wave structure is a carrier of traveling waves. Electromechanical waves on the beam can couple to the circuit by means of the associated electric fields to produce electromagnetic

* Received by the IRE, December 12, 1960; revised manuscript received, February 27, 1961.

† Spencer Lab., Raytheon Co., Burlington, Mass.

[1] D. A. Watkins, "Traveling-wave tube noise figure," PROC. IRE, vol. 40, pp. 65–70; January, 1952.

[2] R. W. Peter, "Low-noise traveling-wave amplifier," *RCA Rev.*, vol. 13, pp. 344–368; September, 1952.

[3] J. R. Pierce, "The wave picture of microwave tubes," *Bell Sys. Tech. J.*, vol. 33, pp. 1343–1372; November, 1954.

waves on the circuit, and vice versa. This coupling is quite analogous to wave coupling between two adjacent circuits.

The electron beam can carry two fundamental kinds of waves, fast waves and slow waves. The phase velocity of a fast wave is greater than the electron velocity, while that of a slow wave is less than the electron velocity. The fast wave can be described as a positive ac energy carrier in the sense that, in order to establish fast-wave modulation on a beam, energy must be transferred to the beam. In this sense, the slow wave is a negative ac energy carrier since, to establish slow-wave modulation, energy must be extracted from the beam. The circuit is obviously a positive ac energy carrier because energy must be fed into the circuit to produce an electromagnetic wave on it.

Coupling between two wave carriers is brought about when the phase velocities of the waves on the carriers are approximately equal. If we adjust the phase velocity of the fast wave to be synchronous with that of the circuit wave, we obtain the kind of coupling which characterizes interaction between any two waves which are positive ac energy carriers. An example of this is the coupling which takes place between two adjacent circuits or transmission lines. This is illustrated in Fig. 1. Signal power entering the coupling region on transmission line no. 2 transfers sinusoidally with distance over to line no. 1. At some distance the power transfer is complete and all of the power is carried on line no. 1. If at this point the lines are separated, the power brought into the coupling region on line no. 2 is taken out of the coupling region on line no. 1. The transfer of power between the fast beam wave and the circuit or transmission-line wave is very similar. This is illustrated in Fig. 2. Under a set of circumstances of particular interest, power is carried by both waves into the coupling region. The transmission line carries signal power and the fast-wave beam noise power. This is the situation depicted in the figure. For a coupling region of just the right length, the fast-wave noise power is completely transferred to the transmission line and is carried away by it, and the signal power is completely transferred to the beam and is carried away by it. In this process no power is amplified. Hence this is not the process which gives amplification in a traveling-wave tube. Nevertheless, this is a very useful process as we shall see in connection with electron-beam parametric amplifiers.

The process which does bring about amplification in a traveling-wave tube involves interaction between the slow wave and the circuit wave. Assume that we adjust the phase velocity of the slow wave to be synchronous with that of the circuit wave. The coupling which takes place is illustrated in Fig. 3. Signal power entering the coupling or interaction region on the transmission line begins to modulate the slow wave. Since slow-wave modulation corresponds to negative power, this shows

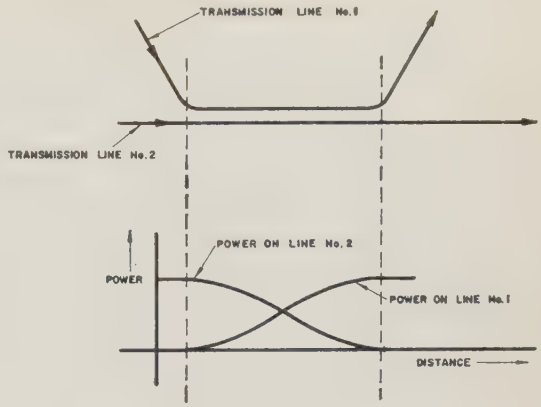


Fig. 1—Coupling between adjacent transmission lines.

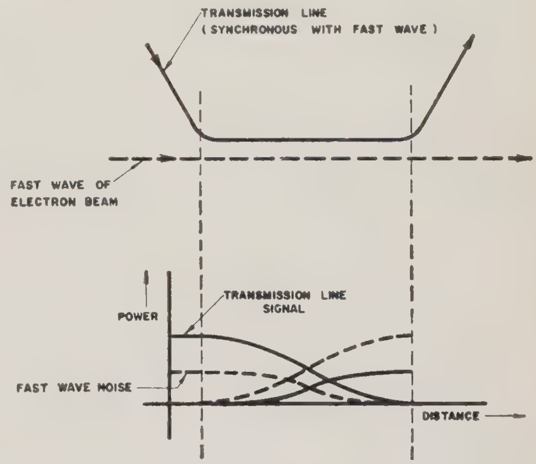


Fig. 2—Coupling between a fast beam wave and a transmission line.

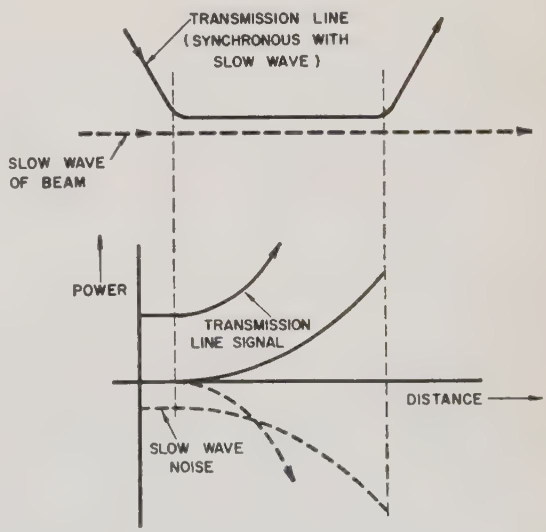


Fig. 3—Coupling between a slow beam wave and a transmission line.

up on the power-vs-distance plot as a line originally tangent to the x axis but extending further and further below it as the modulation builds up. In the coupling previously described between two positive ac energy carriers, we note that the total power on the two lines remains constant with distance (see Figs. 1 and 2.) Only the power distribution between lines changes. This is also true in the present case when the coupling involves one positive ac energy carrier and one negative ac energy carrier. Since the curve representing the slow-wave modulation power extends further and further in the negative direction, the curve representing the transmission-line signal power must extend further and further in the positive direction. It can be shown that both curves increase (one in the negative direction and the other in the positive direction) exponentially with distance. The exponential growth constitutes the amplification.

A similar process of exponential growth on each wave takes place because of the beam noise power which enters on the slow wave. This is also shown in Fig. 3. Note that power due to the transmission-line input signal and also power due to the beam noise appear on both waves at the end of the interaction region. There is no separation of signal and noise which we observed to be possible in the case of fast-wave coupling. The output terminal of the traveling-wave tube is the transmission line as it leaves the interaction region. The output contains power due to beam noise as well as power due to amplification of the input signal.

As might be expected from this description, the beam noise constitutes the main source of noise in the traveling-wave tube. Theory shows that if the beam is accelerated rapidly after it leaves the cathode and if the beam voltage beyond that region is everywhere greater than a few volts, there are severe limitations on the noise reduction which can take place. No practical noise reduction technique or combination of techniques is capable of reducing the noise figure to below about 6 db. This fact gave rise in the past to the erroneous belief that the minimum noise figure under *any condition* was about 6 db.

A new technique has been discovered recently which involves an operation on the beam in the vicinity of the cathode or where the thermal velocity spread is large relative to the average velocity of the electrons.[4–7] Specifically, the beam is made to flow through an extended low-velocity region where the beam voltage is a fraction of a volt. Under these conditions, effective noise reduction has produced noise figures at S band of less than 3 db and at X band of around 4 db.[8–11]

A method of producing the low-velocity region is to use a stacked array of anodes as shown in Fig. 4. The potential of each anode can be adjusted to determine experimentally the optimum condition for lowest noise figure. This technique is operable both with conventional wide-band traveling-wave tubes (whose amplification process was described in connection with Fig. 3) and with narrow-band backward-wave amplifiers (whose amplification process was not described but is similar to that of the traveling-wave tube as far as the noise limitations are concerned).

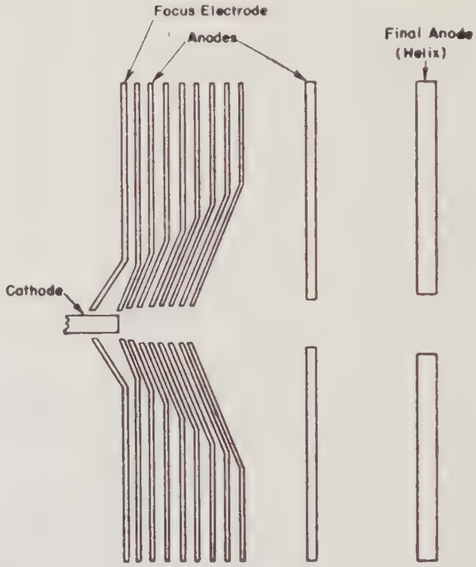


Fig. 4—Stacked array of anodes for noise reduction in beams.

We have used noise figure or noise temperature several times already to describe the noise characteristics of various amplifiers. Although these terms are in rather common usage, it is nevertheless appropriate at this point to pause to emphasize their physical significance. We will be using both noise figure and noise temperature in what follows for purposes of comparison. As illustrated in Fig. 5, the noise figure can be regarded as the

[4] A. W. Siegman, D. A. Watkins, and H. C. Hsieh, "Density-function calculations of noise propagation on an accelerated multi-velocity electron beam," *J. Appl. Phys.*, vol. 28, pp. 1138–1148; October, 1957.

[5] M. R. Currie and D. C. Forster, "Low-noise tunable preamplifiers for microwave receivers," Proc. IRE, vol. 46, pp. 570–579; March, 1958.

[6] M. R. Currie, "A new type of low-noise electron gun for microwave tubes," Proc. IRE, vol. 46, p. 911; May, 1958.

[7] M. Caulton and G. E. St. John, "*S*-band traveling-wave tube with noise figure below 4 db," Proc. IRE, vol. 46, pp. 911–912; May, 1958.

[8] M. R. Currie and D. C. Forster, "New mechanism of noise reduction in electron beams," *J. Appl. Phys.*, vol. 30, pp. 94–103; January, 1959.

[9] B. P. Israelsen, E. W. Kinaman, and D. A. Watkins, "Development of ultra-low-noise traveling-wave amplifiers at Watkins-Johnson Company," *Proc. Symp. on the Application of Low Noise Receivers to Radar and Allied Equipment*, Lincoln Lab., Mass. Inst. Tech., Lexington, Mass., October 24–28, 1960; vol. 2, pp. 257–285, November, 1960.

[10] J. E. Nevins, Jr., "Application of low noise backward wave amplifiers to radar receivers," *Proc. Symp. on the Application of Low Noise Receivers to Radar and Allied Equipment*, Lincoln Lab., Mass. Inst. Tech., Lexington, Mass., October 24–28, 1960; vol. 2, pp. 287–311, November, 1960.

[11] O. Hodowanec and H. J. Wolkstein, "An Ultra-Low-Noise Wide-Band Traveling-Wave Tube," presented at Electron Devices Meeting, Washington, D. C.; October, 1960.

ratio of the total noise power delivered to the load due to all sources to that delivered to the load due to the input source alone. Hence the noise figure is a measure of the amplifier's noisiness relative to the noisiness of the source. The noise figure of an ideal or noiseless amplifier is unity. For any actual or noisy amplifier, we must specify the temperature (or the noisiness) of the input source to uniquely define the noise figure, and this temperature is taken to be room temperature (290°K). If the source is not at room temperature, the ratio given in Fig. 5 is sometimes called the operating noise figure. The physical significance of noise temperature is immediately apparent assuming that the temperature of the input source is adjustable. The noise temperature is that temperature to which the input source must be adjusted to make the operating noise figure precisely 2. Hence when the input source is at the noise temperature, the input source contributes just as much noise to the output of the amplifier as that contributed by sources within the amplifier.

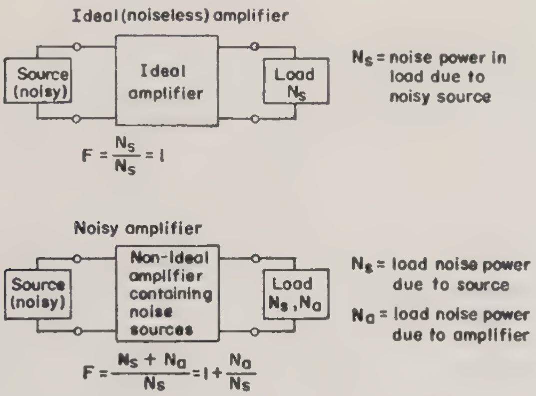


Fig. 5—Physical significance of noise figure.

$$\text{Noise figure} = \frac{\text{total noise power delivered to load}}{\text{noise power delivered to load by source only}}.$$

One of the lowest noise temperatures so far reported for an S-band traveling-wave tube is the 250°K obtained by workers at Watkins-Johnson Company.[9] The broad-band noise temperature for that tube is less than 585°K from 2 to 4 kMc. Comparable results are being obtained at other places, notably at Hughes Aircraft Company and at RCA.[8,10,11] An X-band backward-wave amplifier built at Hughes has a minimum measured noise temperature of 490°K.[10] The noise temperatures for the S-band Watkins-Johnson tube and for the X-band Hughes tube are plotted in Fig. 6 along with noise temperatures measured for other types of amplifiers. The other noise temperature data will be referred to specifically as the other types of amplifiers are discussed later in the paper.

There appears to be no theoretical limit to the noise reduction which can take place in a traveling-wave tube. However, there are inherent practical considerations which give an indication of the ultimate noise characteristics achievable. Although any prediction of future performance is extremely hazardous in this field, a prediction is ventured in Fig. 7. This prediction was published about a year ago by Watkins and the author.[12] It still appears to be valid.

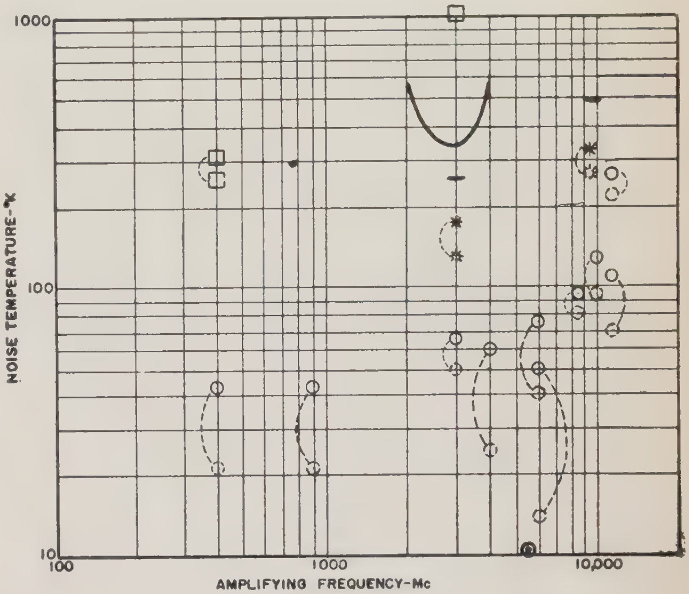


Fig. 6—Present status of measured noise temperatures for various types of amplifiers.
Traveling-wave tubes and backward-wave amplifiers——
Parametric amplifiers, double channel ○
Parametric amplifiers, single channel ✱
Tunnel diodes □
Maser ⊙

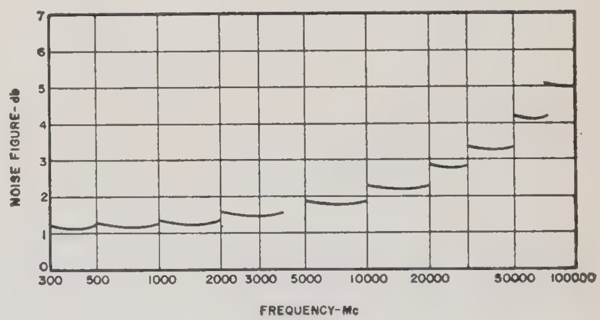


Fig. 7—Predicted noise figures for traveling-wave tubes of the future.

Electron-Beam Parametric Amplifiers

Modulation on a beam can be amplified by a fundamentally different process than is used in the traveling-wave tube. The process is such that amplification of fast beam waves is possible as well as amplification of slow beam waves. We have already seen in connection with Fig. 2 how the beam noise can be coupled out of a fast wave at the same time that a signal is coupled in. If a fast wave in which this has happened is then amplified,

[12] D. A. Watkins and G. Wade, "TWT's and paramps for low-noise reception," *Electronics*, vol. 32, pp. 106–109; December, 1959.

low-noise operation will result. The amplification can be brought about by a parametric process. The motion of an individual electron about its equilibrium position in an electron beam can be pumped parametrically and made to grow.

One way of visualizing this process is to consider a mechanical analog of a single electron in a beam. Fig. 8 shows a pendulum in simple harmonic motion. The pendulum bob is charged positively and its motion is confined to a region occupied by a uniform electric field. Assume that we can control at will the magnitude of the electric field. It is intuitively apparent that we can amplify the pendulum motion by suddenly turning on the field whenever the pendulum bob reaches either of the two positions of maximum excursion and by suddenly turning off the field whenever the pendulum bob swings through its equilibrium position. This constitutes parametric excitation. An analogous process takes place in all parametric amplifiers.

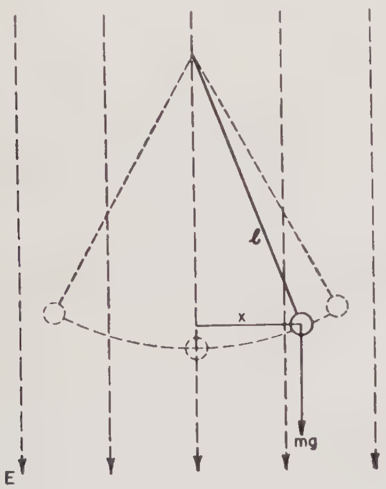


Fig. 8—A pendulum whose motion is pumped parametrically.

The analogy is particularly striking in the case of electron-beam parametric amplifiers. A beam electron which is removed from its equilibrium position oscillates about that position in such a way that its component of motion in a plane containing the beam axis is simple harmonic motion. This is true for both transverse waves of electron motion such as cyclotron waves, and longitudinal waves such as space-charge waves. In this sense, the electron is analogous to a pendulum bob. Since the electron is charged, the analogy extends to a charged pendulum bob. We have already seen how the proper time-varying electric field is capable of pumping the motion of a charged pendulum bob. In analogous fashion the proper time-varying electric field is also capable of pumping the motion of the electron. Since the equilibrium position of the electron travels with the beam, the pumping field, in a sense, must likewise travel with the beam. From considerations of this kind, it is possible to show that, in general, a traveling electric field of the right kind can parametrically amplify the wave motion (for either transverse or longitudinal waves) of all the individual electrons in an electron beam.[13–24] Since the process amplifies the motion of individual electrons, it can be applied to any pattern of beam modulation which produces that motion, whether the motion is associated with fast waves or with slow waves. As we have seen, fast-wave amplification is particularly attractive from the standpoint of low noise. Complete beam noise removal or cancellation is theoretically possible for fast waves. This means that in principle at least, the equivalent beam temperature of a fast-wave amplifier can be reduced to 0°K. In practice, fast cyclotron wave amplifiers have been built with beam temperatures below 25°K. The details of the noise performance thus far achieved will be discussed in a following paragraph.

The basic configuration of an electron-beam parametric amplifier is given in Fig. 9. The input coupler serves two functions. It couples out of the beam the entering fast-wave component of beam noise (thus cooling the beam), and at the same time, it modulates the beam with a new fast wave corresponding to the input signal. In the figure, the coupler is depicted diagrammatically as a circulator because it functions in a fashion similar to the conventional circulator. Power entering the coupling region on the fast wave of the beam (at the point marked 1 in the figure) goes out through the input terminals of the coupler (at the point marked 2). Power entering the coupling region through the input terminals (at the point marked 2) goes out on the fast wave (at the point marked 3). Unlike the conventional circulator, there can be no power transmission from point 3 to point 1, since power cannot be carried upstream by the beam in order to enter at point 3. The

[13] T. J. Bridges, "An electron beam parametric amplifier," Proc. IRE, vol. 46, pp. 494–495; February, 1958.

[14] W. H. Louisell and C. F. Quate, "Parametric amplification of space-charge waves," Proc. IRE, vol. 46, pp. 707–716; April, 1958.

[15] R. Adler, "Parametric amplification of the fast electron wave," Proc. IRE, vol. 46, pp. 1300–1301; June, 1958.

[16] R. Adler, G. Hrbek and G. Wade, "A low-noise electron-beam parametric amplifier," Proc. IRE, vol. 46, pp. 1756–1757; October, 1958.

[17] G. Wade and R. Adler, "A new method for pumping a fast space-charge wave," Proc. IRE, vol. 47, pp. 79–80; January, 1959.

[18] R. Adler, G. Hrbek and G. Wade, "The quadrupole amplifier, a low-noise parametric device," Proc. IRE, vol. 47, pp. 1713–1723; October, 1959.

[19] C. C. Johnson, "Theory of fast-wave parametric amplification," *J. Appl. Phys.*, vol. 31, pp. 338–345; February, 1960.

[20] E. I. Gordon, "A transverse-field traveling-wave tube," Proc. IRE, vol. 48, p. 1158; June, 1960.

[21] D. C. Forster, "Theory of Parametrically-Pumped Longitudinal-Field Electron Beams," California Inst. Tech., Pasadena, Tech. Rept. No. 14; June, 1960.

[22] A. Ashkin, "A Microwave Adler Tube," presented at Internatl. Congress on Microwave Tubes, Munich, Germany; June 7–11, 1960.

[23] A. E. Seigman, "The dc pumped quadrupole amplifier—a wave analysis," Proc. IRE, vol. 48, pp. 1750–1755; October, 1960.

[24] R. W. Gould and C. C. Johnson, "Coupled mode theory of electron-beam parametric amplification," *J. Appl. Phys.*, vol. 32, pp. 248–258; February, 1961.

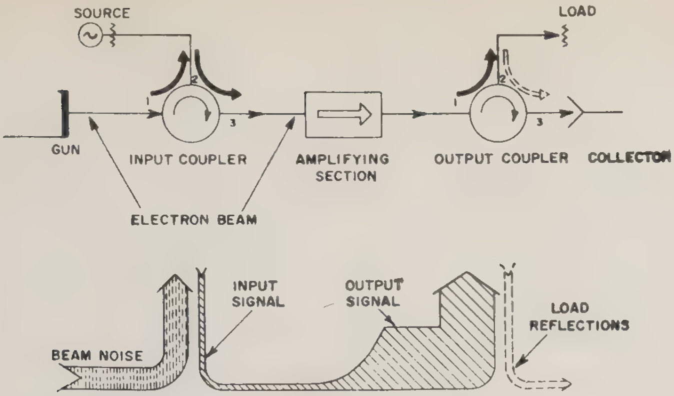


Fig. 9—Basic configuration of an electron-beam parametric amplifier.

output coupler is identical to the input coupler and serves to transmit the amplified signal to the load. Couplers of this kind can be designed to exhibit broadband characteristics.[25]

Noise temperature data for three electron-beam parametric amplifiers is given along with data for other types of amplifiers in Fig. 6. The three electron-beam parametric amplifiers represented are all fast cyclotron wave amplifiers which operate in the near-degenerate mode. The fast space-charge wave amplifiers which have been built have not yet given good low-noise operation. The circles in the figure represent double channel (or radiometer) noise temperatures for electron-beam and solid-state parametric amplifiers operating in the near degenerate mode. The solid-line circles give noise temperatures which include the effect of circulators for the solid-state parametric amplifiers or input couplers (which act as built-in circulators) for the electron-beam parametric amplifiers. The dashed lines extending from the solid circles lead to dashed-line circles which depict the corresponding noise temperatures without the effect of the circulators or input couplers. The three sets of circles pertaining to the electron-beam parametric amplifiers are the ones at 400 Mc, 900 Mc, and 4 kMc, respectively. Some of the lowest noise temperatures yet attained by unrefrigerated amplifiers of any kind have been attained by the electron-beam parametric amplifiers. The amplifier at 4 kMc was built at Bell Laboratories[26] and the other two at Zenith.[27]

Solid-State Parametric Amplifiers

Parametric excitation was illustrated in the previous section in terms of a periodic variation in the downward force acting on a pendulum bob. Excitation of this kind is possible in a wide variety of resonant systems. In general, the excitation is produced by a pump which periodically changes a parameter associated with the energy stored in the system. Depending on the system, such a parameter could be a capacitor, an inductor, a spring constant or, as we have seen, a pendulum force. Parametric devices have been built with vibrating strings, pendulums, electron beams, ferrites and semiconductor material.[28] However different in appearance the various devices may be, they all operate by virtue of a periodic variation in a parameter which controls the energy stored in the system.

In any amplifier there must be an energy source and a means for transferring the energy from the source to the signal to be amplified. In the parametric amplifier the pump is the energy source and the principle of parametric excitation is the means for energy transfer.

It is well known that a semiconductor diode can be made to look like a time-varying capacitance. The pump in this case is simply a source of microwave voltage at the proper frequency applied to the terminals of the diode. The diode exhibits a capacitance whose value is a function of the voltage across it. This property results from a phenomenon which occurs in the depletion layer. Assume that we connect a variable source of voltage across a p-n junction diode so that the positive terminal of the source is attached to the n-type material and the negative terminal, to the p-type material. If we increase the voltage, the depletion layer expands. If we decrease the voltage, the depletion layer contracts. In a sense, the depletion layer is like the space between the plates of a capacitor. As the depletion layer widens, the corresponding capacitance decreases; as the depletion layer becomes narrower, the capacitance increases. The depletion layer can expand and contract at microwave frequency rates. This means that by supplying microwave voltage across the diode terminals, we can produce a capacitance which varies at a microwave frequency. Hence we have the necessary ingredient for microwave parametric excitation and therefore for microwave amplification.

A diagram of a basic circuit for a parametric amplifier which uses a variable capacitance is shown in Fig. 10. The two circles depict two resonant cavities which are coupled in series with the capacitance. The diagram has been simplified to the extent that it does not indicate how the pump voltage is supplied to produce the capacitance variation. Ordinarily this voltage is coupled across the diode by means of a third resonant cavity. The pump frequency is designated by ω_p, and the two resonant frequencies are ω_1 and ω_2, respectively. An in-

[25] J. Herman and J. Litton, Jr., "Characteristic features of a broad band beam parametric amplifier," *Proc. Symp. on the Application of Low-Noise Receivers to Radar and Allied Equipment*, Lincoln Lab., Mass. Inst. Tech., Lexington, Mass., October 24–28, 1960; vol. 2, pp. 331–346; November, 1960.

[26] A. Ashkin, Bell Telephone Labs., Inc., Murray Hill, N. J. (private communication).

[27] R. Adler, Zenith Radio Corp., Chicago, Ill. (private communication).

[28] For excellent bibliographies on parametric devices see W. W. Mumford, "List of selected references on parametric amplifiers (to July, 1959)," Proc. IRE, vol. 48, pp. 850–853; May, 1960, and E. Mount and B. Begg, "Parametric devices and masers: an annotated bibliography," IRE Trans. on Microwave Theory and Techniques, vol. MTT-8, pp. 222–224; March, 1960.

put signal at ω_1 is fed into the tank on the left—the tank which is resonant to that frequency. An amplified version of that signal is extracted from the output of the same tank. An idler signal at ω_2 is generated in the second tank and can be extracted from it if desired. For single-channel operation the idler signal is not extracted and, if ω_1 and ω_2 are separated sufficiently, the first tank can provide the filtering in both the input and the output of the amplifier, and thereby eliminate interference at the idler frequency. Often in the case of double-channel operation ω_1 and ω_2 are so nearly equal that only one resonant tank is needed. Signals at both the amplifying and the idler frequencies then appear simultaneously within the single tank. Under these conditions the input and the output couplings will pass both the amplifying and the idler frequency channels.

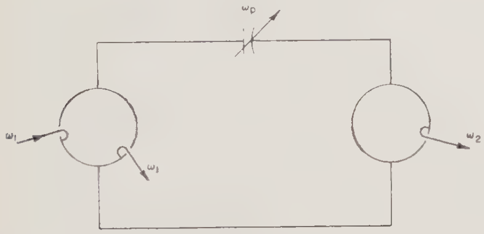


Fig. 10—A basic configuration for a variable-capacitance parametric amplifier. $\omega_p=\omega_1+\omega_2$.

The amplifier as described above is of the negative conductance type. Because of the parametric excitation, the variable capacitance and the idler tank appear to couple a negative conductance into the input tank. The gain expression therefore has the familiar form exhibited by all regenerative amplifiers.

The most significant source of noise in this amplifier is thermal noise from resistive elements in the circuit. The capacitance variation can be obtained with operation at convenient temperatures such as room temperature, or at lower temperatures if lower noise is desired. From the standpoint of noise, the lower the temperature, the better. Thermal noise is due to disorganized molecular agitation. Since the degree of agitation is proportional to the absolute temperature, the noise energy also is proportional to the absolute temperature. The lowest noise operation obtained from any nonmaser amplifier has been obtained from a refrigerated parametric amplifier of this type.

The diode itself gives rise to a fundamental limitation on the noise performance of the amplifier. In addition to providing the variable capacitance depicted in Fig. 10, all diodes have a constant resistance in series with the capacitance. The resistance is associated with the loss in the bulk semiconductor, the contacts and the leads. Several authors have published performance analyses which take into account the series resistance.[29–34] An important result of these analyses is the prediction of optimum pump frequency and minimum noise figure.

Noise temperature data for nine semiconductor-diode parametric amplifiers are shown in Fig. 6. As previously stated, the solid-line and dashed-line circles represent double-channel noise temperatures with and without circulators, respectively. The two sets of stars give similar information about single-channel noise temperatures. The data were taken by workers at Raytheon, Hughes, Bell Laboratories and Westinghouse as follows:[35]

Raytheon—
- Single-channel noise temperature at 3 kMc.
- Double-channel noise temperature at 10 kMc.

Hughes—
- Double-channel noise temperature at 3 kMc.
- Double-channel noise temperature at 8.6 kMc.

Bell Laboratories—
- Double-channel noise temperature at 6 kMc.
- Double-channel noise temperature at 6 kMc. (refrigerated)
- Double-channel noise temperature at 11.5 kMc.
- Double-channel noise temperature at 11.5 kMc. (refrigerated)

Westinghouse—single-channel noise temperature at 9.5 kMc. (refrigerated)

The lowest noise temperature yet achieved by any parametric amplifier excluding the circulator is the 15°K at 6 kMc measured by M. Uenohara of Bell Laboratories with an amplifier using a gallium arsenide diode refrigerated to 90°K.

As previously indicated, microwave parametric amplifiers have been built with ferrite material also. However, at present, the development of ferrite parametric amplifiers is much less advanced than that of diode parametric amplifiers or electron-beam parametric am-

[29] P. Penfield, Jr., "Noise in negative-resistance amplifiers," IRE Trans. on Circuit Theory, vol. CT-7, pp. 166–170; June, 1960.

[30] R. C. Knechtli and R. D. Weglein, "Low-noise parametric amplifier," Proc. IRE, vol. 48, pp. 1218–1226; July, 1960.

[31] K. E. Mortenson, "Parametric diode figure of merit and optimization," *J. Appl. Phys.*, vol. 31, pp. 1207–1212; July, 1960.

[32] K. L. Kotzebue, "Optimum noise performance of parametric amplifiers," Proc. IRE, vol. 48, pp. 1324–1325; July, 1960.

[33] P. Penfield, Jr., "Fundamental Limits of Varactor Parametric Amplifiers," Microwave Associates, Inc., Burlington, Mass., Tech. Rept.; August, 1960.

[34] H. H. Grimm and C. R. Boyd, Jr., "Optimum design of semiconductor low-noise circuits for an *L*-band radar system," *Proc. Symp. on the Application of Low-Noise Receivers to Radar and Allied Equipment*, Lincoln Lab., Mass. Inst. Tech., Lexington, Mass., October 24–28, 1960; vol. 2, pp. 173–191, November, 1960.

[35] I. Goldstein, Raytheon Co., Bedford Lab., Bedford, Mass.; R. D. Weglein, Hughes Res. Labs., Malibu, Calif.; M. Uenohara, Bell Telephone Labs., Murray Hill, N. J.; and T. Hollis, Westinghouse Electric Corp., Baltimore, Md. (private communications).

plifiers. Although the ferrite amplifiers may well be important in the future, at present they are not competitive in terms of low noise. One of the best amplifiers of this kind constructed so far has given 20-db gain at 4.5 kMc with a noise figure of 12 db.[36]

Tunnel-Diode Amplifiers

The tunnel-diode (or Esaki-diode) amplifier is the most recent negative-resistance microwave device.[37] The fact that such a diode can provide negative resistance is apparent from its characteristic curve—a curve similar to that of the old screen-grid tetrode or dynatron. Typical characteristic curves (current versus voltage) for the Esaki diode and for the conventional semiconductor diode are shown in Fig. 11. The amount of resistance represented on such a curve is inversely proportional to the steepness of the slope. A portion of the curve for the Esaki diode slopes downward from left to right. In that region, an increase in voltage causes a decrease in current, and hence, the resistance there is negative.

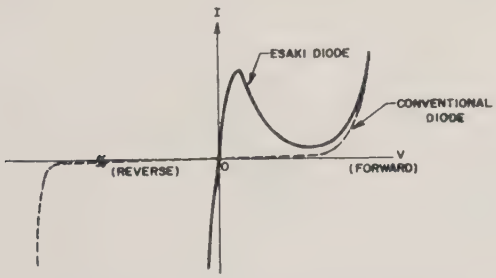


Fig. 11—Characteristic curves for the Esaki diode and for the conventional semiconductor diode.

The negative-resistance effect is due to tunneling of charge carriers through a potential barrier. Classical physics cannot explain tunneling. The explanation requires a quantum mechanical viewpoint. The tunneling in an Esaki diode is similar to the tunneling which takes place in high-field emission from the surface of a cathode. In both cases, the carriers have insufficient thermal energy to go over the top of the potential barrier in producing current flow. However, they are able to tunnel through the barrier. Quantum mechanics tells us that it is impossible to clearly define particles as small as electrons. They behave as waves of probability. The density of such a wave is greatest in the center and gradually tails off with distance. If an electron is close to a very thin potential barrier, its wave of probability extends beyond the barrier; and, if there is an unfilled energy state for it to occupy, an appreciable probability exists for its being beyond the barrier. Hence, when a great many charge carriers are located on one side of a thin potential barrier, an appreciable number must be on the other side of the barrier. Since the carriers could not have climbed over the barrier, we regard them as having tunneled through it.

It is well known that semiconductor material, such as silicon or germanium, becomes a fairly good conductor through the insertion of impurities. The *p-n* junction of the conventional semiconductor diode is formed at the plane which separates two regions of opposite-type impurities. The familiar energy-band picture for the vicinity of the *p-n* junction in such a diode is shown in the upper half of Fig. 12. The mechanism for current flow is diffusion. For example, when forward bias is applied, electrons in the conduction band on the *n* side cross the junction by diffusing over the potential hill to the *p* side without leaving their allowed band. This mechanism gives the characteristic shown in Fig. 11 for the conventional diode.

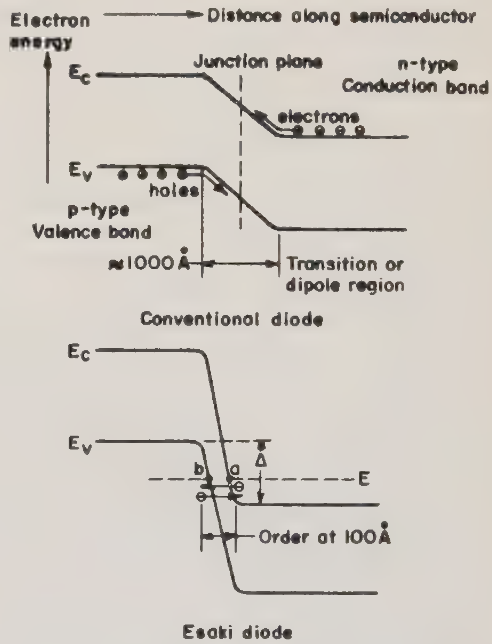


Fig. 12—Energy-band picture for the conventional diode and for the Esaki diode.

Dr. Leo Esaki discovered that a large increase in the level of the impurity concentration produces a significant change in the energy-band picture. He experimented with an impurity concentration of the order of a million times higher than that in ordinary diodes. Under this condition, the energy-band picture is the one shown in the lower half of Fig. 12. In the *p*-type material, the upper portion of the valence band is practically cleared, and in the *n*-type material, the lower portion of the conduction band is practically filled.

[36] R. Denton, "A Ferrite Parametric Amplifier," presented at PGMTT National Symp., San Diego, Calif.; May 9–11, 1960.

[37] For a bibliography on tunnel-diode amplifiers, see J. J. Tiemann, "Shot noise in tunnel diode amplifiers," Proc. IRE, vol. 48, pp. 1418–1423; August, 1960.

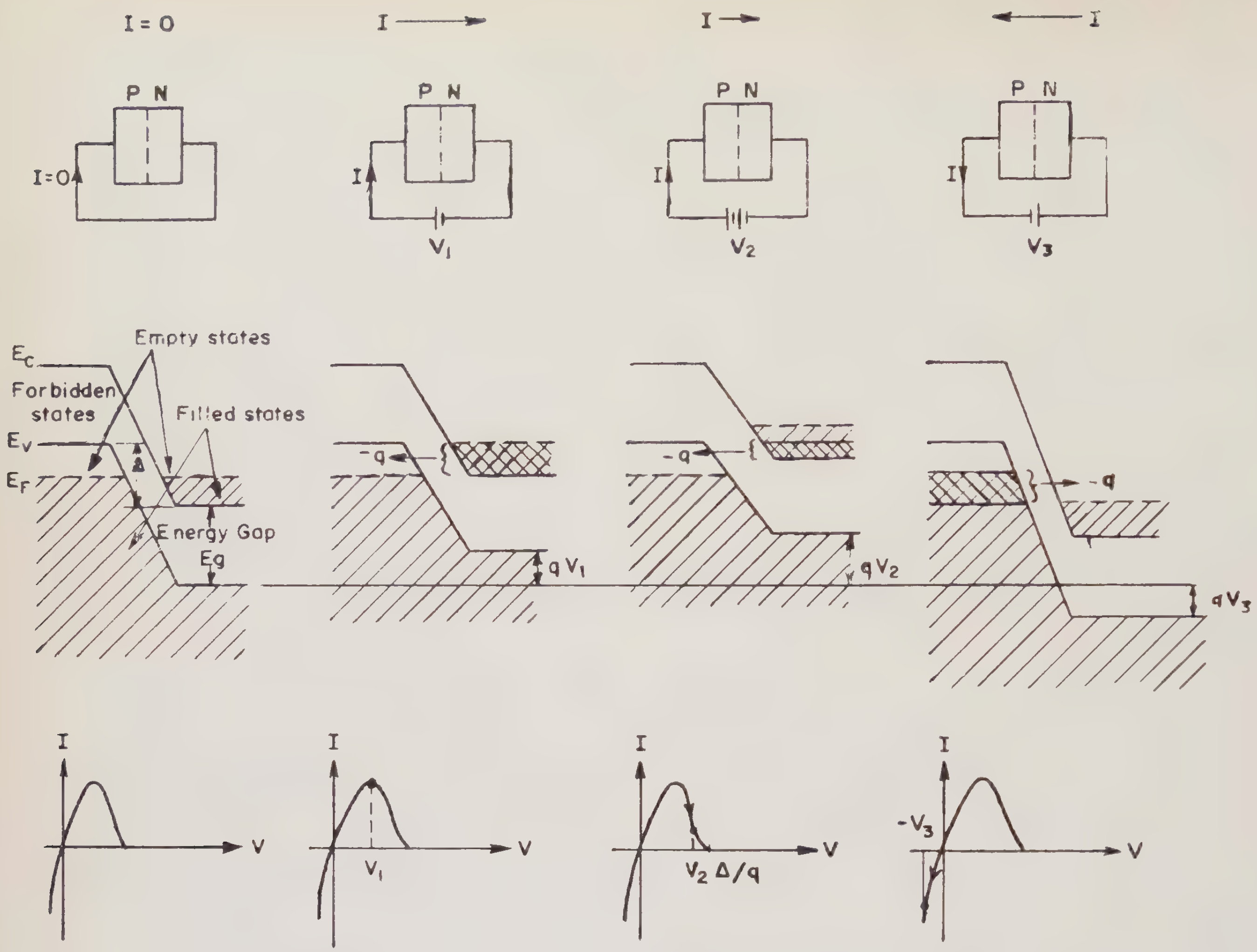


Fig. 13—Origin of the tunnel currents and the negative-resistance characteristic.

Hence, the Fermi level (marked E in the figure) is actually inside the main bands as indicated. The energy-band picture shows that some energy levels in the valence band of the p-type material are in line with some levels in the conduction band of the n-type material. In addition, the depletion layer, or transition region, is much narrower than in the conventional diode.

The conditions for tunneling exist. For example, an electron in the lower part of the conduction band of the n-type material is no longer opposite a forbidden band in the p-type material, but is separated by only a short distance (through the narrow forbidden band of the transition region) from a permissible state in the p-type material. If the permissible state is unoccupied, tunneling can take place.

We can now understand the reason for the negative-resistance characteristic of the Esaki diode. For simplicity, assume that the temperature is absolute zero. All energy states below the Fermi level are filled, and all above it are empty. Fig. 13 illustrates the situation. For zero applied voltage (the condition represented at the extreme left), there are no filled states opposite empty states on either side, and so there is no current. A small forward voltage raises the energy levels in the n-type material relative to those in the p-type material and produces a condition in which filled states are opposite empty states. Tunneling occurs, as indicated, by the arrow in the energy-band diagram second from the left, and current flows in the forward direction. As the forward voltage is increased beyond a certain point, the number of filled states opposite empty states falls off, and the tunneling decreases. Consequently, the increase in forward voltage results in a decrease in the forward current. The negative-resistance region arises from this situation. The current, under actual conditions, does not quite dip to zero as Fig. 13 (with the idealizations assumed) indicates. The actual characteristic is the one previously shown in Fig. 11.

A major source of noise is shot noise in the dc current which must flow as the amplifier operates.[37,38] This is illustrated in Fig. 14. Tunnel-diode amplifiers are usually operated with a bias corresponding very nearly to that of maximum negative slope on the characteristic

[38] P. Penfield, Jr., "Noise performance of tunnel-diode amplifiers," Proc. IRE (Correspondence), vol. 48, pp. 1478–1479; August, 1960.

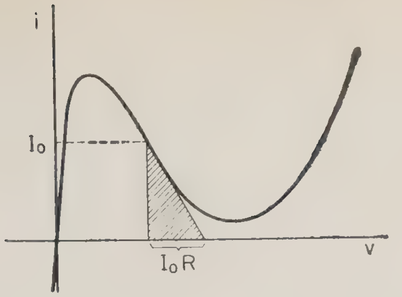


Fig. 14—The minimum theoretical noise figure for a tunnel-diode amplifier, assuming that all thermal noise sources are negligible.

In limit as $R_s \to 0$,

$$F_{\min} = 1 + \frac{e}{2kT_0} I_0 R = 1 + 20 I_0 R.$$

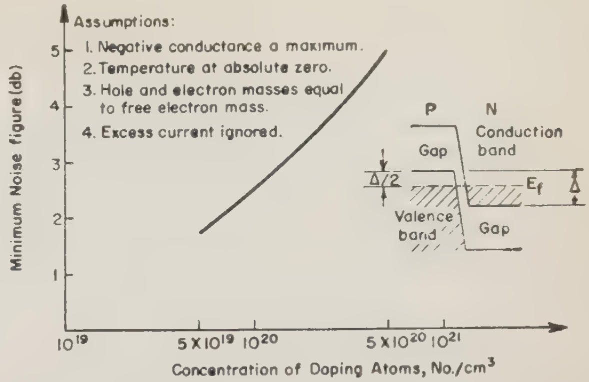


Fig. 15—Minimum theoretical noise figure as a function of concentration of doping atoms.

curve. The shot-noise term in the noise-figure expression is given in Fig. 14, and is proportional to the distance along the voltage axis between the bias voltage and the intersection of the tangent to the characteristic curve at the operating point. From the shape of the curve, it is intuitively obvious that the minimum noise figure is obtained when the bias is close to (but slightly higher than) that corresponding to maximum negative slope.

R. Pucel of Raytheon[39] has calculated noise-figure data as a function of impurity, or doping concentration, assuming only the shot-noise source. These data are plotted in Fig. 15. The calculations are based on the simplified tunneling model originally suggested by Esaki.[40] A number of the assumptions used are indicated in the figure. The noise-figure curve decreases with decreasing concentration. The lower the concentration the better, as far as noise behavior is concerned. However, there is an obvious limit to how far the concentration can be decreased. As we have discussed, the tunneling, and hence the negative resistance, is brought about by virtue of high concentration. If we make the concentration too low, we no longer have a tunnel-diode. The lower portion of the curve shown in the figure corresponds to a reasonable lower limit for the concentration.

The above calculations seem to be consistent with experimental results. Noise-temperature data for two tunnel-diode amplifiers are shown in Fig. 6. The solid-line and the dashed-line squares represent the noise temperatures with and without circulators respectively. The data were obtained by workers at RCA.[41]

Masers

The maser appears to be the ultimate in low-noise amplification at microwave frequencies.[42] Strictly speaking, the word maser applies only to microwave frequencies. It stands for *m*icrowave *a*mplification by *s*timulated *e*mission of *r*adiation. Related devices at higher frequencies have been designated properly as irasers (for *i*nf*ra*red frequencies) and lasers (for *l*ight frequencies).

As Weber[43] has pointed out, a number of types of amplifiers, including free-electron vacuum-tube amplifiers, operate through stimulated emission of radiation. However, for the purposes of this discussion, let us consider only the three-level molecular amplifier.

Fig. 16 illustrates, schematically, how the maser works. Assume a system with three energy levels E_1, E_2, and E_3, and assume that transitions can occur between each level and either of the other two. The figure indicates the population of molecules corresponding to each level. In thermal equilibrium, the population follows a Boltzmann distribution as shown. An intense RF pumping field at the frequency f_p can induce transitions between levels 1 and 3 and can equalize the populations of these levels. Under these conditions, the population of level 3 may be greater than that of level 2. If a weak signal at frequency f_a is applied, more transitions are produced from level 3 to level 2 than from level 2 to level 3. Hence, there is a net emission of radiation from the system at the frequency f_a. More energy leaves the system at that frequency than enters it. This is the basic mechanism for amplification in a maser.

An inherent source of unavoidable noise in masers is due to spontaneous emission. A molecule in level 3 can drop spontaneously to level 2 and emit a quantum of radiation. This radiation is not coherent with any stimulating signal. It is produced in a purely random process, and consequently has the character of noise. The product of the frequency and the ratio of Planck's constant to Boltzmann's constant gives approximately the contribution of this source to the noise temperature.[44]

[39] Private communication.

[40] L. Esaki, "New phenomenon in narrow germanium *p-n* junctions," *Phys. Rev.*, vol. 109, pp. 603–604; January, 1958.

[41] K. K. N. Chang, RCA, Princeton, N. J. (private communication).

[42] For a bibliography on masers, see E. Mount and B. Begg, "Parametric devices and masers: an annotated bibliography," IRE Trans. on Microwave Theory and Techniques, vol. MMT-8, pp. 222–243; March, 1960.

[43] J. Weber, "Masers," *Rev. Mod. Physics*, vol. 31, pp. 681–710; July, 1959.

[44] J. Weber, "Maser noise considerations," *Phys. Rev.*, vol. 108, p. 537; November, 1957.

At microwave frequencies the contribution is small. However, at frequencies above the millimeter wave region, the contribution is substantial. For example, at 10 microns the noise temperature due to this source is greater than 1000°K. At light frequencies, it is greater than 10,000°K. Fig. 17 shows a plot of the approximate expression for this noise temperature as a function of frequency.

Because of spontaneous emission, neither lasers nor irasers can be nearly as sensitive as masers (since masers operate at much lower frequencies). For example, consider an iraser being used to detect blackbody radiation at 10 microns. The radiator would have to be at a temperature above 1000°K to contribute as much power to the output of the iraser as does the iraser's own spontaneous emission. Photon counters, which will be discussed briefly in the next section, are not subject to spontaneous emission and consequently, in principle, have lower detection limits.

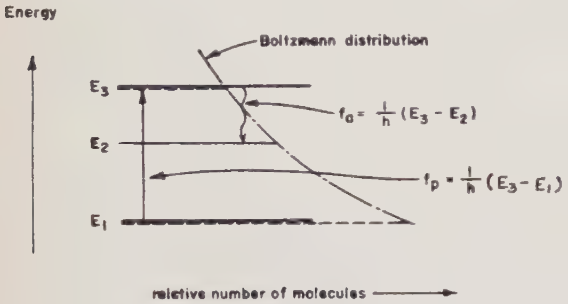


Fig. 16—Population of molecules in the energy levels of a three-level maser.

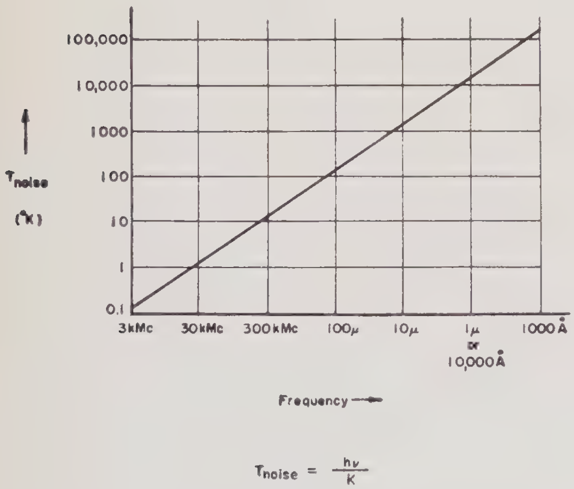


Fig. 17—The approximate noise temperature due to spontaneous emission.

The lowest noise temperatures yet measured for any microwave amplifiers are the noise temperatures for masers. Fig. 18 gives the performance data for a well-designed traveling-wave maser. The data were obtained by DeGrasse, Hogg, and colleagues at Bell Laboratories.[45] If we exclude the noise contributions from the sky, the antenna loss, and the antenna waveguide loss, we have for the maser proper a noise temperature of only 10.5°K. This is the lowest value plotted in Fig. 6. Since the traveling-wave maser does not need a circulator, this point should be compared to the solid-line circles, stars, and squares in the figure. If we leave out the effect of the input connection or maser coax, we have left for the intrinsic maser only about 2°K.

It is interesting to compare further the properties of the various types of low-noise amplifiers we have discussed. Fig. 19 presents this comparison. Note that the values indicated in the various columns for any one type of amplifier do not necessarily all apply to a single amplifier of that particular type. For example, no single amplifier of the maser class (masers, irasers and lasers) can have a measured noise temperature (intrinsic) of less than 5°K and at the same time operate at optical frequencies.

TW Maser Performance
(De Grasse, Hogg, Ohm, Scovil)

Signal frequency	5.65 kMc
Bandwidth	25 Mc
(Tunable over 230 Mc)	
Gain	35 db
Noise contributions:	
Sky	2.5°K
Sidelobes	2.0°K
Antenna loss	1.5°K
Waveguide loss	2.0°K
Maser coax	8.5°K
Intrinsic maser noise	2.0°K
Over-all system noise temperature	18.5°K

Fig. 18—Performance data for a traveling-waver maser.

Device	Measured Noise Temperature	Practical Upper Frequency Limit	Bandwidth
Diode Parametric Amplifiers	10–20°K	Order of 50 kMc	Order of several hundred Mc
Electronic Beam Parametric Amplifiers	20–30°K	Order of 10 kMc	Order of several hundred Mc
Masers, Irasers, Lasers	<5°K	Optical frequencies and higher	Order of several hundred kMc
Tunnel Diodes	150–300°K	Order of 30–100 kMc	Order of 1 kMc

Fig. 19—Comparison of properties of low-noise amplifiers.

[45] R. W. DeGrasse, formerly of Bell Telephone Labs., Murray Hill, N. J., presently at Microwave Electronics Corp., Palo Alto, Calif. (private communication).

Photon Counters and Photosensitive Detectors

As was discussed, spontaneous emission limits the sensitivity of maser-type detectors and is particularly important at higher frequencies. Weber has pointed out that spontaneous emission is also present in conventional vacuum-tube amplifiers.[44] However, there is a type of detector in which spontaneous emission does not play a part. It is possible to construct a detector which employs particles in their ground states. Various photosensitive IR detectors are of this type, as are the photon counters proposed by Bloembergen[46] and Weber.[43]

Consider Bloembergen's scheme. Fig. 20 illustrates how it works. It involves a material with three energy levels. The material is refrigerated so that all the molecules are in the ground state (level 1). If a pumping signal is applied at the frequency corresponding to transitions between levels 2 and 3, no effect is produced since level 2 is unoccupied. However, incoming photons at the frequency corresponding to transitions between levels 1 and 2 can produce a detectable effect. The incoming photons can raise particles to level 2, where the pump can raise them further to level 3. The resulting spontaneous emission from the upper level can be detected by optical means.

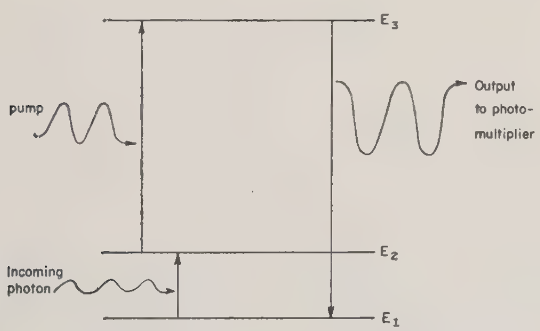


Fig. 20—Photon counter of the type proposed by Bloembergen.

Note that in Bloembergen's proposed device, there is no output of any kind (noise or otherwise) until a photon arrives. Initially, all the particles are in the ground state. No devices of precisely this kind have yet been built. However, photosensitive IR detectors have been built which, in principle, can also operate with particles initially in the ground state. Absorbed infrared radiation excites electrons from the valence band of a semiconductor to the conducting state. The change in conduction is then detected. In the maser (and in all other amplifiers having particles which are initially excited), there is a noise output at all times. In the ideal photon counter and photosensitive detector, there is no output until a photon appears. The presence of any output is evidence of some input.[47] Spontaneous emission is not present.

Conclusion

We have considered a number of new techniques for low-noise detection at centimeter wavelengths and at shorter wavelengths. Within the last ten years, the attainable noise temperatures have been reduced by two and three orders of magnitude. Among the devices which are currently competitive or which show future promise as far as low-noise performance is concerned, are traveling-wave tubes, parametric amplifiers, tunnel diodes, masers, photon counters and photosensitive detectors. Each of these devices has certain advantages and disadvantages for any particular application. Each has its place. Each will be with us and serving us in important ways for a number of years to come.

[46] N. Bloembergen, "Solid state infrared quantum counters," *Phys. Rev. Lett.*, vol. 2, pp. 84–85; February, 1959.

[47] Note that phase information is not retained between input and output.

General Relativity for the Experimentalist*

ROBERT L. FORWARD†

Summary—**Einstein's general theory of relativity is broken down and simplified under limitations usually satisfied in experimentally realizable situations. Following the work of Møller,[1] an analogy between electromagnetism and gravitation is presented which allows calculation of various gravitational forces by considering the equivalent electromagnetic problem. A number of examples are included. Tensor formulation is not used except in the Appendix, where justification for the analogies is given.**

Introduction

We SHALL assume that all gravitational effects are correctly described by Einstein's general theory of relativity.[2] Suppose we want to find the behavior of a system under the influence of gravitational and other forces in the proper general relativistic manner. We must first determine all the mass and energy in both the system being investigated and in the sources of the fields. Using these in a prescribed manner, we calculate the ten components of the energy-momentum tensor. Next, we put the energy-momentum tensor into Einstein's equations and solve these ten nonlinear differential equations for the ten components of the metric tensor. Then the metric tensor is substituted into the curvilinear equation of motion and it is solved to determine how the system moves.

Because of the difficulty of this process, there exist only a few solutions of Einstein's equation which are of experimental interest in that they describe some physically observable effect of general relativity. The process is so specialized and so difficult that it is practically impossible to attempt solution of a problem unless months of study on the specialized terminology, procedures, and conventions of the general relativity theorist have been completed. The most disappointing aspect is that in most cases, after struggling through the calculations, it will only be found that the effect calculated is too small to be observed.

Fortunately, it is possible to bypass this complicated procedure. By applying reasonable limitations to the systems involved, Einstein's equations can be linearized. These linearized equations can be examined and handled separately. One equation gives us the gravitational scalar potential; three others give a quantity which is similar to the magnetic vector potential; and the remainder, which describe the remaining properties of the gravitational field, can be handled by assuming that they represent a curvature to the space.

With these analogies, it is possible for anyone who has had electromagnetic theory to study a situation of experimental interest and to calculate the effects to be expected with sufficient accuracy to determine whether they warrant further study.[3]

There have been numerous books, papers,[4–6] articles,[7] and even advertisements[8] that have developed or mentioned a gravitational analog to the electric and magnetic fields. Some are based on classical ideas, some on Mach's principle, and others on Einstein's general theory of relativity. None based on Einstein's work, however, have been at the same time complete, rigorous, and free of tensors.

Analogies to Electricity and Magnetism

In Einstein's general theory of relativity there exist gravitational analogies to the electric and magnetic fields. We are already familiar with the analogy between the electric field of a charge and the gravitational field of mass. It is well known that the analogy breaks down almost immediately, because there are two kinds of charge and only one kind of mass and because two particles with the same type of charge repel, whereas two particles with the same type of mass attract. Nevertheless, if we are cognizant of these distinctions, we can still apply the analogy and obtain useful results.

In the Appendix it is shown that Einstein's equations not only can be made to show this well-known analogy between the electric field and the gravitational field, but they can also give a gravitational analog to the magnetic field. It has no name, but since it will be mentioned, we shall coin one and call it the "protational field," as it usually arises from the rotation of a mass. This "protational field" has the dimensions of angular velocity (radians/sec) and is closely related to coriolis-type forces which arise from the principle of general relativity.

When the analogy is carried out and all the constants

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[1] C. Møller, "The Theory of Relativity," Oxford University Press, London; 1952.

[2] A. Einstein, "The foundation of the general theory of relativity," *Ann. Phys.*, vol. 49, pp. 769–822; May 11, 1916. See also, A. Einstein, "The Principle of Relativity," Dover Publications, Inc., New York, N. Y.; 1923.

[3] If anyone using this approximate method comes upon a previously uncalculated effect that shows promise of being large enough to be observed, the author will be glad to repeat the calculations using the proper tensor formulation to ensure that the result was not produced by the approximations involved in simplifying the theory.

[4] D. W. Sciama, "On the origin of inertia," *Mon. Not. Roy. Astr. Soc.*, vol. 113, pp. 34–42; January, 1953.

[5] W. D. White, "Electromagnetic analogs for the gravitational fields in the vicinity of a satellite," Proc. IRE, vol. 46, pp. 920–922; May, 1958.

[6] W. Davidson, "General relativity and Mach's principle," *Mon. Not. Roy. Astr. Soc.*, vol. 117, pp. 212–224; February, 1957.

[7] G. P. Field, "Two source field theory," essay submitted to Gravitational Research Foundation, New Boston, N. H. Some of the terminology and symbols used were adopted from this essay.

[8] W. D. White, A.I.L. Advertising Monographs, Proc. IRE, vol. 46, p. 4A; November and December, 1958 and January, 1959.

are evaluated, we obtain an isomorphism between the gravitational and the electromagnetic quantities (see Table I).

Now, if we have a certain mass distribution and flow, all that is necessary is to find a similar charge and current distribution in electromagnetic texts, such as that of Smythe.[9] We then use the formulas derived for the electric and magnetic fields and make the substitutions in the electromagnetic formulas to obtain the gravitational formulas. Since we are using the linearized theory, superposition is valid, and fields for more complex bodies can be built up from the superposition of the fields of the parts.

Once we have calculated the fields generated by the mass density and currents, we can calculate the forces on a particle of mass m by a force equation which is analogous to the electromagnetic force equation,

$$\begin{aligned} \boldsymbol{F} &= -m\nabla\chi - m\frac{\partial \boldsymbol{K}}{\partial t} + m\boldsymbol{v}\times(\nabla\times\boldsymbol{K}) \\ &= m\boldsymbol{G} + m(\boldsymbol{v}\times\boldsymbol{P}). \end{aligned} \tag{1}$$

If the test body is spinning and has an angular momentum of $\boldsymbol{L}$, then the torque on it due to a "protational field" $\boldsymbol{P}$ will be by analogy

$$\boldsymbol{N} = \tfrac{1}{2}\boldsymbol{L}\times\boldsymbol{P}. \tag{2}$$

It should be emphasized that the previous discussion is approximate and is presented merely to provide a simple tool with which to make estimates. In deriving this analogy between some of the gravitational forces and the static and induction fields of electromagnetism, the following assumptions, among others, have been made:

1) The mass densities are normal (no dwarf stars).
2) All motions are much slower than the speed of light. (Often special relativistic effects will hide general relativistic effects.)
3) The kinetic or potential energy of all the bodies being considered is much smaller than their mass energy.
4) The fields are always weak enough so that superposition is valid.
5) The distances between objects is not so large that we have to take retardation into account. (This condition can be ignored when we have a stationary problem where the fields have already been prescribed and are not changing with time.)

To show how this analogy can be used, let us calculate a few simple examples.

Force between Two Masses

Suppose we have two particles with total masses M_1 and M_2. Then if we want to calculate the gravitational field due to mass M_1, we write down the electric field for a particle with charge Q_1:

$$\boldsymbol{E} = \frac{Q_1}{4\pi\epsilon r^2}\hat{\boldsymbol{r}}. \tag{3}$$

Next we transform all quantities to the equivalent gravitational quantities and get

$$\boldsymbol{G} = \frac{-M_1}{4\pi\gamma r^2}\hat{\boldsymbol{r}}. \tag{4}$$

Then the force on a particle of mass M_2 in the gravitational field of M_1 is

$$\boldsymbol{F} = M_2\boldsymbol{G} = \frac{-M_1M_2}{4\pi\gamma r^2}\hat{\boldsymbol{r}}, \tag{5}$$

and if we transform γ into more familiar gravitational units, we get

$$\boldsymbol{F} = \frac{-GM_1M_2}{r^2}\hat{\boldsymbol{r}}, \tag{6}$$

which is merely Newton's law.

[9] W. R. Smythe, "Static and Dynamic Electricity," McGraw-Hill Book Co., Inc., New York, N. Y.; 1950.

TABLE I

	EM Symbol		Gravitational Symbol	Value or Definition
Force Vector	$-\boldsymbol{E}$	$\rightarrow$	$\boldsymbol{G}$	$=-\nabla\chi-\dfrac{\partial \boldsymbol{K}}{\partial t}$
Solenoidal Force Vector	$-\boldsymbol{B}$	$\rightarrow$	$\boldsymbol{P}$	$=\nabla\times\boldsymbol{K}$
Scalar Potential	$-\phi$	$\rightarrow$	χ	$\approx-\dfrac{1}{4\pi\gamma}\displaystyle\int_V\frac{\mu}{r}dV$
Vector Potential	$-\boldsymbol{A}$	$\rightarrow$	$\boldsymbol{K}$	$\approx-\dfrac{\eta}{4\pi}\displaystyle\int_V\frac{\mu\boldsymbol{v}}{r}dV$
Source Density	ρ	$\rightarrow$	μ	$=\dfrac{dM}{dV}$
Source Quantity	Q	$\rightarrow$	M	$=\displaystyle\int_V\mu dV$
Current Density	$\boldsymbol{j}$	$\rightarrow$	$\boldsymbol{p}$	$=\mu\boldsymbol{v}$
Current Quantity	I	$\rightarrow$	T	$=\dfrac{dM}{dt}=\displaystyle\int_s\boldsymbol{p}\cdot\boldsymbol{n}dS$
Moment	$\boldsymbol{M}$	$\rightarrow$	$\dfrac{1}{2}\boldsymbol{L}$	$=\dfrac{1}{2}I\omega$
Capacitivity of Space	ϵ	$\rightarrow$	γ	$=\dfrac{1}{4\pi G}=1.19\times10^9\,\dfrac{\text{kg-sec}^2}{\text{m}^3}$
Permeability of Space	μ	$\rightarrow$	η	$=\dfrac{16\pi G}{c^2}=3.73\times10^{-26}\,\dfrac{\text{m}}{\text{kg}}$

Pinch Effect between Two Pipes

This example is included primarily to show why the gravitational equivalent of the magnetic field has never been observed.

Suppose that molten metal is flowing through two parallel pipes with spacing $d=0.1$ meter. To find their interaction due to the mass currents T_1 and T_2, we look at the equivalent magnetic case of two wires with a current I. There will exist a pinch effect caused by a force-per-unit length of

$$\frac{F}{L} = \frac{\mu I_1 I_2}{2\pi d}. \tag{7}$$

We then transform all the quantities to get the equivalent gravitational relation:

$$\frac{F}{L} = \frac{\eta T_1 T_2}{2\pi d}. \tag{8}$$

Let us be overly generous and assume that molten iridium is used so that each meter of pipe will contain about 200 kg and we can obtain a flow velocity of a kilometer per second. Then the mass current in each pipe will be

$$T_1 = T_2 = Mv = 2 \times 10^5 \text{ kg/sec.}$$

The force between the two pipes due to the "protational" pinch effect will be about 2×10^{-15} newtons per meter of pipe.

If, for comparison, we also look at the gravitational attraction for the same two pipes and use Newton's law, we get a force of about 3×10^{-4} newtons per meter of pipe, so that the forces due to the pinch effect are hidden by the gravitational effect, which itself is not usually observable. This is quite discouraging, but it is the usual result of gravitational calculations. Furthermore, this estimate of the effect has saved us the labor of calculating it by using tensor quantities.

Satellites of Spinning Bodies

We shall now estimate the effect of the earth's rotation on an artificial satellite. Any effects will probably be hidden by the perturbations induced by inhomogeneous mass distributions, atmospheric friction, and even the light pressure from the sun; however, we shall calculate them anyway.

First we need to know the "protational field" of the earth. From Smythe[9] we find an expression for the external magnetic field produced by a ring current I at a latitude $\theta=\alpha$ on a spherical shell of radius R. By transforming the magnetic quantities into the equivalent gravitational quantities, we obtain an expression for the "protational field" of a rotating massive ring with a mass current T:

$$P_\theta = \frac{-\eta T \sin\alpha}{2R} \sum_{n=1}^{\infty} \frac{1}{(n+1)} \left(\frac{R}{r}\right)^{n+2} \cdot P_n{}^1(\cos\alpha) P_n{}^1(\cos\theta). \tag{9}$$

Since it is assumed that superposition is valid, we can construct the "protational field" of a solid spinning body by integrating over the volume:

$$P_\theta = \frac{-\eta\omega\sin\theta}{8\pi r^3} \int_V [\mu(\alpha, R) R^2 \sin^2\alpha] R^2 \sin\alpha\, d\alpha\, d\phi\, dR + \text{higher multipoles.} \tag{10}$$

Since $R \sin\alpha$ is the distance from the axis of rotation to the mass element, we see that the integral is merely the moment of inertia I of the body. Thus, in general, the "protational field" of any rotating body is approximately

$$P_\theta \approx \frac{-\eta I\omega \sin\theta}{8\pi r^3}. \tag{11}$$

Similarly, it can be shown that

$$P_r \approx -\frac{\eta I\omega\cos\theta}{4\pi r^3}. \tag{12}$$

Now that we know the "protational field" of a spinning body, such as the earth, we can calculate the effect of this field on a satellite.

A satellite in a polar orbit around a spinning body would experience a perturbing force due to the radial component of the "protational field." (See Fig. 1.) Neglecting space curvature, we find that this force would cause a precession of the orbit of an amount

$$\Omega = \frac{\boldsymbol{N}}{\boldsymbol{L}} = \frac{F_\phi r}{mrv_\theta} = P_r. \tag{13}$$

Averaging the effect over the whole orbit gives

$$\Omega = \frac{-2}{\pi} \frac{\eta I\omega}{4\pi r^3}. \tag{14}$$

Now, if we substitute numbers for the case of a satellite in a polar orbit around the earth,

$$I = 8.11 \times 10^{37} \text{ kg-m}^2$$
$$r = 7.4 \times 10^6 \text{ meters}$$
$$\eta = 3.73 \times 10^{-26} \text{ m/kg}$$
$$\omega = 7.29 \times 10^{-5} \text{ rad/sec,}$$

we find a precession of the orbit of 5.5×10^{-14} rad/sec. This is equivalent to a period of rotation of the orbital

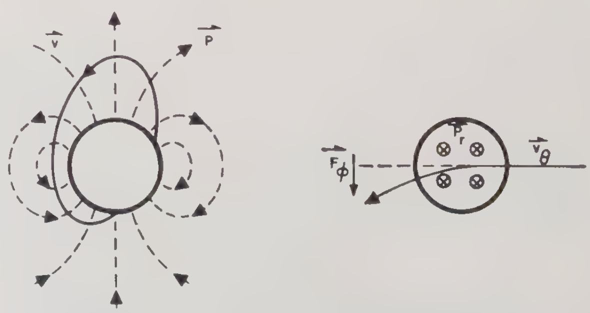

Fig. 1—Effect of earth's "protational field" on a satellite in a polar orbit.

plane of 27 billion years, which is indeed too slow to be seen.

A satellite in an equatorial orbit would experience a radial force due to the tangential component of the "protational field." This force would be completely hidden by the radial Newtonian gravitational force and would be observable only if the earth were completely symmetric and we knew its mass to the nth decimal place. If the satellite is spinning, however, then the spin axis of the satellite itself will experience a torque,

$$N = \tfrac{1}{2}L \times P \tag{15}$$

due to spin-spin interaction. If we ignore space curvature effects, this torque will cause a precession of the spin axis by an amount,

$$\Omega = \frac{N}{L} = \frac{P}{2} \sin \beta. \tag{16}$$

For a satellite in an equatorial orbit of radius r around a spinning body with angular momentum $I\omega$, we get

$$\Omega = \frac{P_\theta}{2} = \frac{-\eta I \omega}{16\pi r^3} = -\frac{GI\omega}{c^2 r^3}\cdot \tag{17}$$

The problem of the precession of a spinning satellite has been rigorously calculated in the proper relativistic manner by L. I. Schiff.[10] His equation for the precession rate was

$$\Omega = \frac{3GM}{2c^2 r}\,\omega_{\text{orbit}} - \frac{GI}{c^2 r^3}\,\omega_{\text{spin}}. \tag{18}$$

The first term, the largest, is due to the effects of space curvature, which we neglected. This term is independent of the spin of either the satellite or the planet. The second term is the one we calculated. If we substitute numbers for the case of a spinning satellite in equatorial orbit around the earth,

$$\begin{aligned} r &\approx 7.4 \times 10^6 \text{ meters} \\ \omega_{\text{spin}} &= 7.29 \times 10^{-5} \text{ rad/sec} \\ \omega_{\text{orbit}} &\approx 10^{-3} \text{ rad/sec} \\ I &= 8.11 \times 10^{37} \text{ km-m}^2 \\ M &= 5.98 \times 10^{24} \text{ kg,} \end{aligned}$$

we get

$$\Omega = (9.0 - 0.11) \times 10^{-13} \text{ rad/sec.} \tag{19}$$

Thus we see that the spin interaction term is a small percentage of the space curvature term.

Analogy to Electromagnetic Radiation

Since we have derived an analogy between electromagnetism and gravitation, we might naively suppose that this analogy also would hold for electromagnetic radiation. We might consider writing the Maxwell relations, transforming them to the equivalent gravitational relations, and then solving them to get the wave equation. If we do this using the analogy for the electric and magnetic fields, we shall find that the equations describe a wave with a propagation constant of half the speed of light, since in electromagnetic theory $1/\sqrt{\mu\epsilon} = c$ and in "graviprotational" theory $1/\sqrt{\gamma\eta} = c/2$. We are reasonably sure, however, that the velocity of propagation of gravitational energy will be the same as the speed of light, since the value obtained by Einstein for the rotation of the perihelion of Mercury depends upon this value. Thus we have another indication that our analogy is not perfect, but will give order-of-magnitude estimates only.

Despite the failure of this analogy, it is possible by taking more terms into account to show that Einstein's equation contains the proper wave equation. In the Appendix we obtain

$$\Delta^2 \phi_{\alpha\beta} - \frac{1}{c^2} \frac{\partial^2 \phi_{\alpha\beta}}{\partial t^2} = 0, \tag{20}$$

where $\phi_{\alpha\beta}$ is a quantity representing the gravitational potential of an accelerated mass. The interpretation of this equation is that an accelerated mass will emit gravitational waves which travel with the velocity of light.

No one has ever observed pure gravitational radiation, and from the examples at the end of this section, we shall see why. The observance of interaction in the induction or near-field zone of an accelerated mass is quite another matter. With a sensitive torsion balance, Cavendish[11] observed the attraction of one mass by another and measured the value of the gravitational constant (Fig. 2). If we swing the large masses back and

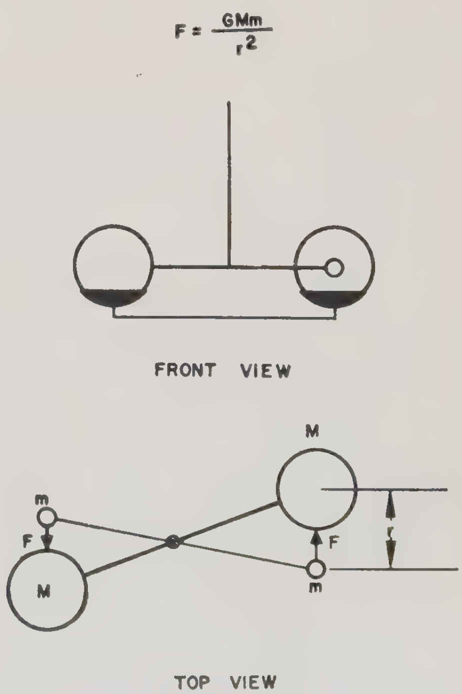

Fig. 2—Cavendish's experiment.

[10] L. I. Schiff, "Possible new experimental test of general relativity, theory," *Phys. Rev. Lett.*, vol. 4, pp. 215–217; March 1, 1960.

[11] H. Cavendish, *Phil. Trans. Roy. Soc.*, vol. 17, p. 469; 1798.

forth with the same period as the natural period of the torsion pendulum, it is easy to see that oscillations will build up in the pendulum and resonant absorption will occur in the near zone of the large masses.

It is well known that the induction fields are conservative and that if there were no resonant absorption, there would be no losses due to the near field. In electromagnetic theory, if we examine the fields at a large distance from the field generator, then the near field becomes negligible and all that remains is the radiation field, which is not conservative. A radiation field carries away energy, and the oscillations in the generator damp out as a result of the radiation losses. It is this gravitational equivalent of the radiation field that has never been observed, either directly or by radiation damping of a mechanically accelerating system.

In Landau and Lifshitz,[12] the wave equation for the gravitational potential is solved and transformed from a four-dimensional relationship into a temporal-spatial relationship. The general solution is

$$\phi_{ab} = \frac{2G}{c^4 r}\int_V \mu x_a x_b dV. \qquad a, b = 1, 2, 3. \tag{21}$$

By calculating the energy in a plane wave at large distances from the source and averaging over all directions, the total energy emitted per unit time in all directions by the accelerated mass is given by

$$-\frac{dE}{dt} = \frac{G}{45c^5}\sum_{a=1}^{3}\sum_{b=1}^{3}(\dddot{Q}_{ab})^2, \tag{22}$$

where

$$Q_{ab} = \int_V \mu(3x_a x_b - \delta_{ab} r^2)dV$$

is the mass quadrupole moment of the mass source.

Note that energy will not be radiated unless the source has an accelerated mass quadrupole moment. Thus, gravitational waves must be quadrupole radiation or higher multipole radiation. There is no dipole gravitational radiation; this is easily seen by physical arguments. Suppose that we grasp a charged particle and shake it, *i.e.*, accelerate it. Since it is the only moving charge in the area, it emits electromagnetic dipole radiation. Now suppose that we hold a particle with mass and shake it. As we rapidly accelerate the small particle in one direction, our large body, in order to conserve momentum, will slowly accelerate in the opposite direction. Because the "charge-to-mass" ratio in gravitation is unity, the two accelerated bodies will always radiate the same amount of dipole radiation, but they will be out of phase and therefore the dipole radiation will cancel.

Quantum Relations for Gravitational Radiation

Gravitational radiation never has been observed and general relativity has not been quantized; therefore, the following statements are only educated theoretical guesses.

1) Gravitational radiation is quantized. The elementary quanta have been named gravitons.
2) The spin of a graviton is 2. This is basically because gravitational radiation can only be of the quadrupole type. Photons, being dipole radiation, have a spin of 1.
3) The velocity of a graviton is the same as the velocity of a photon and is related to the frequency and wavelength in the same way:

$$c = f\lambda = 3 \times 10^8 \text{ m/sec}.$$

4) The energy and momentum of a graviton depends upon Planck's constant in the same way as does a photon:

$$E = hf$$
$$p = h/\lambda.$$

Gravitational Radiation from a Spinning Dumbbell

The simplest quadrupole mass source for the calculation of gravitational-radiation energy emission is two equal masses rotating about their center of mass.

We first calculate the mass quadrupole moment with respect to a fixed co-ordinate system. Let us assume that

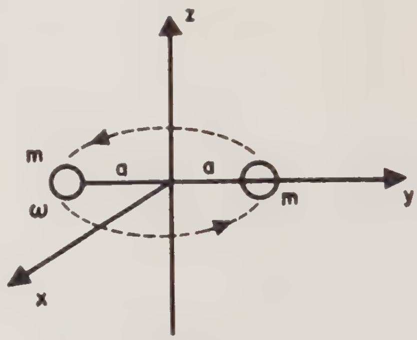

Fig. 3—Spinning dumbbell.

the spin axis is in the z-direction (Fig. 3), then

$$\begin{aligned} Q_{11} &= \int_V \mu(3x^2 - r^2)dV \\ &= 2m(2a^2\cos^2\omega t - a^2\sin^2\omega t) \\ Q_{21} &= Q_{12} = 3\int_V \mu xy dV = 6ma^2\cos\omega t\sin\omega t \\ Q_{22} &= \int_V \mu(3y^2 - r^2)dV \\ &= 2m(2a^2\sin^2\omega t - a^2\cos^2\omega t) \\ Q_{33} &= -2m(a^2\sin^2\omega t + a^2\cos^2\omega t) \\ &= -2ma^2, \end{aligned} \tag{23}$$

and all other quadrupole moments are zero.

Secondly, we calculate the third derivative, and we are left with only the x, y components,

[12] L. Landau and E. Lifshitz, "The Classical Theory of Fields," Addison-Wesley Publishing Company, Inc., Reading, Mass.; 1951.

$$\dddot{Q}_{11} = -\dddot{Q}_{22} = 24(2ma^2)\omega^3 \sin \omega t \cos \omega t$$
$$\dddot{Q}_{12} = \dddot{Q}_{21} = -12(2ma^2)\omega^3(\cos^2 \omega t - \sin^2 \omega t), \quad (24)$$

since the first derivative of the z component is zero.

The power radiated is then

$$-\frac{dE}{dt} = \frac{G}{45c^5}[\dddot{Q}_{11}{}^2 + \dddot{Q}_{12}{}^2 + \dddot{Q}_{22}{}^2 + \dddot{Q}_{21}{}^2] = \frac{8GI^2\omega^6}{5c^5}, \quad (25)$$

where $I = 2\ ma^2$ is the moment of inertia of the source.

Now we must substitute numbers into (25). For a 1-meter dumbbell weighing 1 metric ton and spinning at about 10,000 rpm, conditions which no known material can withstand, the power radiated is only

$$-\frac{dE}{dt} = 4.5 \times 10^{-33} \text{ watts.} \quad (26)$$

With numbers such as these, it is not surprising that this field has been of little interest to experimentalists.

From the exponents of I^2 and ω^6 in (25), it seems desirable, at first glance, to work with a higher rotational speed, even if it means that less mass could be used. However, we would find, when the strength of the material is considered, that it is more advantageous to lower the rotational speed and to use a greater mass. The ultimate in this procedure is represented by a rotating double star system. The rotational speed could be on the order of 1 month $\omega = 10^{-4}$ rad/sec, and the moment of inertia would then be roughly $I = ma^2 = 10^{30}$ kg$\times 10^{12}$ m^2; thus, the power radiated from a binary star is about 10^7 watts. This appears to be a large amount of power, but it would take 10^{10} billion years for the system to damp out as a result of radiation losses.

Space Curvature

The previous analogies have shown us how to calculate the *forces* exerted on a body as a result of the gravitational scalar and vector potential. However, if we are interested in the path of the particle under the influence of the forces, we encounter nonlinearities. It should be noted that we have not yet calculated the usual general relativistic effects, such as the precession of the perihelion of Mercury or the bending of light rays. This is because these effects are not a result of the gravitational-field components which have analogies in electromagnetic theory. Also, when the precession of a spinning satellite was calculated using the electromagnetic analogy, we obtained a result which was smaller than the space curvature effect which we neglected.

It will probably be true in most cases where the primary mass is large and the motions of the bodies have small velocities and accelerations that the only observable perturbations will be a result of the spatial tensor components of the gravitational field, which have no analogy in electromagnetic theory. The closest analogy which may be made is that these components of the gravitational field can be represented by assuming that the mass of the object somehow causes the space to be slightly curved. Then the concept of motion in a flat space under the influence of tensor forces can be replaced by the concept of forceless motion in a curved space.

In classical theory, once we have calculated the forces, we can solve the equation of motion:

$$\frac{d\boldsymbol{p}}{dt} = m\ddot{\boldsymbol{x}} = \boldsymbol{F}(\boldsymbol{x}) \quad (27)$$

for $\boldsymbol{x} = \boldsymbol{x}(t)$. This equation is valid in flat space; however, if we want to include these other gravitational forces by assuming a curved space, then the ordinary rules of differentiation will not hold and we must use covarient differentiation. The usual equation of motion is really just the flat space approximation of the curvilinear equation of motion:

$$\frac{D_c p_a}{Dt} = \frac{dp_a}{dt} - \frac{1}{2}\sum_{b=1}^{3}\sum_{c=1}^{3}\left(\frac{\partial g_{bc}}{\partial x^a}\right)v^b p^c = F_a$$
$$a = 1, 2, 3, \quad (28)$$

where g_{bc} is the three-dimensional metric tensor, and the forces due to the gravitational vector and scalar potential are contained in $\boldsymbol{F}$, v^b is the velocity in the b direction and p^c is the momentum in the c direction.

This procedure should not be too unusual since we know that ordinary rules of differentiation hold only for flat cartesian coordinate systems. In curved coordinate systems, such as cylindrical or spherical systems, we have to use more general rules of differentiation. For instance, the general equation for the divergence of a vector is

$$\nabla\cdot\boldsymbol{A} = \frac{1}{(g_{11}g_{22}g_{33})^{1/2}}\left[\frac{\partial\sqrt{g_{22}g_{33}}\,A_1}{\partial x^1} + \frac{\partial\sqrt{g_{11}g_{33}}\,A_2}{\partial x^2} + \frac{\partial\sqrt{g_{11}g_{22}}\,A_3}{\partial x^3}\right] \quad (29)$$

if the metric tensor is

$$g_{ab} = \begin{pmatrix} g_{11} & 0 & 0 \\ 0 & g_{22} & 0 \\ 0 & 0 & g_{33} \end{pmatrix}$$

And since the spatial metric tensor for cartesian coordinates $x^1 = x$, $x^2 = y$, $x^3 = z$ is

$$g_{ab} = \begin{pmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{pmatrix},$$

we get the familiar relation

$$\nabla\cdot\boldsymbol{A} = \frac{\partial A_1}{\partial x} + \frac{\partial A_2}{\partial y} + \frac{\partial A_3}{\partial z}\cdot \quad (30)$$

For a spherical coordinate system, $x^1 = r$, $x^2 = \theta$, $x^3 = \phi$, however, the spatial metric tensor is not constant, but

depends upon the position in the space,

$$g_{ab} = \begin{bmatrix} 1 & 0 & 0 \\ 0 & r^2 & 0 \\ 0 & 0 & r^2 \sin^2 \theta \end{bmatrix},$$

and we find that calculating the divergence in this curved space is no longer a simple procedure:

$$\nabla \cdot \boldsymbol{A} = \frac{1}{r^2} \frac{\partial(r^2 A_1)}{\partial r} + \frac{1}{r \sin \theta} \frac{\partial(\sin \theta A_2)}{\partial \theta} + \frac{1}{r \sin \theta} \frac{\partial A_3}{\partial \phi}. \quad (31)$$

We also encounter similar problems in simple Newtonian mechanics when spherical or cylindrical coordinate systems are used. For instance, if we want to calculate the motion of a satellite, the equations of motion in cartesian coordinates (let $x^3 = z = 0$) are

$$m\ddot{x} = -\frac{GMm}{(x^2 + y^2)^{3/2}} x$$

$$m\ddot{y} = -\frac{GMm}{(x^2 + y^2)^{3/2}} y, \quad (32)$$

and in spherical coordinates (let $x^2 = \theta = \pi/2$) they are

$$m\ddot{r} - \frac{m}{2} \frac{\partial g_{33}}{\partial r} \dot{\phi}^2 = m\ddot{r} - mr\dot{\phi}^2 = -\frac{GMm}{r^2} \quad (33)$$

$$\frac{d}{dt}(r^2 \dot{\phi}) = 0. \quad (34)$$

It can be seen that we have a term in (33) introduced by the metric tensor of the coordinate system; this is the familiar centrifugal "force." Thus, we can say either that the centrifugal "force" is a real force and that the coordinate space is flat and then use the ordinary equation of motion,

$$m\ddot{r} = \boldsymbol{F} = \boldsymbol{F}_{\text{centrifugal}} + \boldsymbol{F}_{\text{gravitational}},$$

or we can say that there is only one force, that due to the gravitational attraction. Since we are working in a curved coordinate space, however, we must use the curvilinear equation of motion:

$$m\ddot{r} - mr\dot{\phi}^2 = \boldsymbol{F} = \boldsymbol{F}_{\text{gravitational}}.$$

Metric Tensor Outside a Massive Body

We have shown that a curved coordinate system can be interpreted as a force. Now we shall attempt to explain how the remaining components of the gravitational force can be interpreted as a curved space. In practically every imaginable case, the spatial components of the metric tensor will be determined by the scalar potential and special relativity. For the sake of simplicity, we shall continue to ignore special relativity (although it should be taken into account for quantitatively correct results).

The metric tensor for a region of space with a scalar potential χ is given by the Schwarzschild metric tensor, which is a slight modification of the spherical coordinate metric tensor

$$g_{ab} = \begin{vmatrix} \left(1 + \frac{2\chi}{c^2}\right)^{-1} & 0 & 0 \\ 0 & r^2 & 0 \\ 0 & 0 & r^2 \sin^2 \theta \end{vmatrix}, \quad (35)$$

In all cases of experimental interest $1 >> 2\chi/c^2$, and therefore,

$$\left(1 + \frac{2\chi}{c^2}\right)^{\pm n} \approx \left(1 \mp n \frac{2\chi}{c^2}\right) \quad (36)$$

The metric tensor can be written for the common coordinate systems as follows: for spherical coordinate systems, $x^1 = r$, $x^2 = \theta$, $x^3 = \phi$,

$$g_{ab} = \begin{vmatrix} 1 - \frac{2\chi}{c^2} & 0 & 0 \\ 0 & r^2 & 0 \\ 0 & 0 & r^2 \sin^2 \theta \end{vmatrix}; \quad (36)$$

for cartesian coordinate systems, $x^1 = x$, $x^2 = y$, $x^3 = z$,

$$g_{ab} = \begin{vmatrix} 1 - \frac{2\chi}{c^2} & 0 & 0 \\ 0 & 1 - \frac{2\chi}{c^2} & 0 \\ 0 & 0 & 1 - \frac{2\chi}{c^2} \end{vmatrix}; \quad (37)$$

and for cylindrical coordinate systems, $x^1 = \rho$, $x^2 = \phi$, $x^3 = z$,

$$g_{ab} = \begin{vmatrix} 1 - \frac{2\chi}{c^2} & 0 & 0 \\ 0 & \rho^2 & 0 \\ 0 & 0 & 1 - \frac{2\chi}{c^2} \end{vmatrix}. \quad (38)$$

Relativistic Equation of Motion

From the Appendix, we can now write the proper equation of motion which uses this space curvature analogy:

$$\sum_{b=1}^{3} \frac{1}{\Gamma} \frac{d}{dt} (\Gamma g_{ab} \dot{x}^b) = \frac{1}{2} \sum_{b=1}^{3} \sum_{c=1}^{3} \frac{\partial g_{bc}}{\partial x^a} \dot{x}^b \dot{x}^c - \frac{\partial \chi}{\partial x^a} - \frac{\partial K_a}{\partial t} + \sum_{b=1}^{3} \left[\frac{\partial K_b}{\partial x^a} - \frac{\partial K_a}{\partial x^b} \right] v^b, \quad (39)$$

where g_{ab} is one of the metric tensors taken from the previous section, $a = 1, 2, 3$, and

$$\Gamma = \left\{\left[\left(1+\frac{2\chi}{c^2}\right)^{1/2} - \frac{K_a v^a}{c^2}\right]^2 - \frac{v^2}{c^2}\right\}^{-1/2}$$

$$\approx \left\{1+\frac{2\chi}{c^2} - \frac{v^2}{c^2}\right\}^{-1/2}$$

comes from the special and general relativistic corrections to the mass. Note that if $\chi = 0 = K_a$, then we get just the usual special relativistic correction to the mass:

$$m = m_0\Gamma = \frac{m_0}{\sqrt{1-\frac{v^2}{c^2}}}. \tag{40}$$

Now that we have the equation of motion, all that is necessary is to calculate the gravitational scalar and vector potential by analogy to electromagnetism, calculate Γ and g_{ab} and put them in the equation, and then solve the equation by the method of successive approximations.

Measurements in a Curved Space

We should be familiar with the problems of operating in a curved space, since we live on a two-dimensional curved space—the earth. A straight line on the earth is the great circle route, because it is the shortest distance between two points in that two-dimensional space. If we make a triangle with these "straight lines," we find that the sum of the angles can range from π to 5π, depending upon the size of the triangle.

A more fundamental experiment is the parallel translation of a vector. Suppose we are on a flat surface and we place a test vector at one corner of a triangle. Then very carefully keeping the angle between the vector and the appropriate side of the triangle constant, we traverse the perimeter of the triangle and return to the starting point (see Fig. 4). The test vector obviously returns to the starting point with the initial orientation. Now try this same simple experiment with a vector moving about on a spherical triangle, as shown in Fig. 5. It will be obvious even to a flatlander inhabiting the surface of the sphere that the vector has rotated through an angle α as a result of its parallel translation around a closed path in the two-dimensional curved space. The size of the angle α will depend upon the amount of curvature of the space and the size of the triangle.

Effect of Space Curvature on a Satellite

It was pointed out in the section on satellites of spinning bodies that a proper solution of Einstein's equations for the precession of the spin axis of a satellite resulted in two terms. The smaller was a result of the gravitational equivalent of the magnetic field. The larger term was produced by the other components of the gravitational field. The precession of the spin axis of a satellite in a curved space cannot be calculated easily, but by using an analogy to two-dimensional curved space, we can see the qualitative reason for this precession.

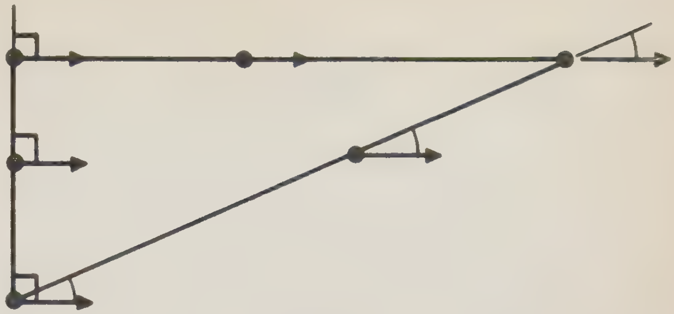

Fig. 4—Parallel translation in flat space.

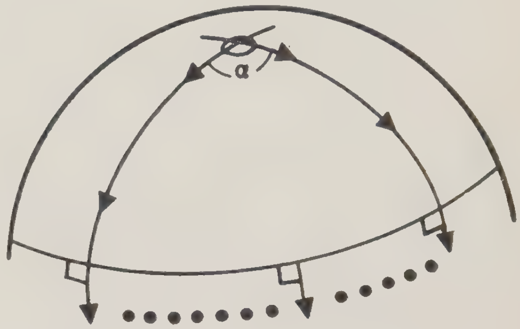

Fig. 5—Parallel translation on a sphere.

We said that these remaining components of the gravitational field of a mass in three-dimensional space could be examined by assuming a slightly curved space with a metric

$$g_{ab} = \begin{pmatrix} \dfrac{1}{1-\dfrac{2GM}{c^2 r}} & 0 & 0 \\ 0 & r^2 & 0 \\ 0 & 0 & r^2\sin^2\theta \end{pmatrix}, \tag{41}$$

where we have taken the scalar potential for a spherical mass $\chi = -GM/r$. Notice that if we are far away from the perturbing mass, then our space is flat again.

Now suppose that we look at a flatlander living on the edge of a two-dimensional massive circle in a two-dimensional flat universe (Fig. 6). If we assume that Einstein's law of gravitation holds in this two-dimensional space, then there will be a tensor gravitational field with which to contend. It will be easier for the flatlanders to ignore these tensor fields and assume that the massive circle warps the local area of his flat space and modifies his two-dimensional metric tensor slightly:

$$g_{ab} = \begin{pmatrix} \dfrac{1}{1-\dfrac{2GM}{c^2 r}} & 0 \\ 0 & r^2 \end{pmatrix}.$$

If a two-dimensional rocket is sent up and a two-dimensional satellite is put into orbit, then the flatlanders will notice that an axis of the satellite will rotate because of the local curvature of the space (Fig. 7). If the orbit is sufficiently distant, the extra contribution to the metric

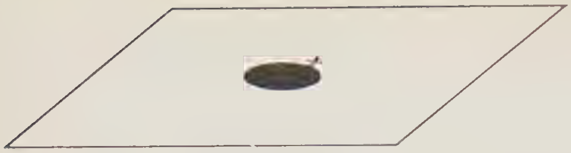

Fig. 6—Flat two-dimensional space.

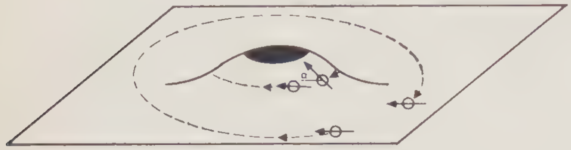

Fig. 7—Two-dimensional locally curved space.

tensor will be small and the satellite will travel through flat space with essentially no rotation.

If this picture is extended to three dimensions and the proper calculations are performed, we obtain the precession due to space curvature:

$$\Omega = \frac{3GM}{c^2 r}\,\omega_{\text{orbit}}$$

Calculation of the Orbit of Mercury

We start from the curvilinear equation of motion, where the only force is that due to the gradient of the scalar gravitational potential:

$$\sum_{b=1}^{3}\frac{1}{\Gamma}\frac{d}{dt}\left(\Gamma g_{ab}\frac{dx^b}{dt}\right) = \frac{1}{2}\sum_{b=1}^{3}\sum_{c=1}^{3}\frac{\partial g_{bc}}{\partial x^a}v^b v^c - \frac{\partial\chi}{\partial x^a} \qquad (42)$$

where $a=1, 2, 3$, and

$$\Gamma = \left(1 + \frac{2\chi}{c^2} - \frac{v^2}{c^2}\right)^{-1/2}.$$

Keeping only the nonvanishing terms, the three equations are:

$$\frac{1}{\Gamma}\frac{d}{dt}(\Gamma g_{11}\dot{r}) = \frac{1}{2}\left[\frac{\partial g_{11}}{\partial r}\dot{r}^2 + \frac{\partial g_{22}}{\partial r}\dot{\theta}^2 + \frac{\partial g_{33}}{\partial r}\dot{\phi}^2\right] - \frac{\partial\chi}{\partial r}$$

$$\frac{1}{\Gamma}\frac{d}{dt}(\Gamma g_{22}\dot{\theta}) = \frac{1}{2}\left[\frac{\partial g_{33}}{\partial\theta}\dot{\phi}^2\right]$$

$$\frac{1}{\Gamma}\frac{d}{dt}(\Gamma g_{33}\dot{\phi}) = 0. \qquad (43)$$

The scalar potential, due to the sun, and its gradient are

$$\chi = -\frac{GM}{r}, \qquad \frac{\partial\chi}{\partial r} = \frac{GM}{r^2}.$$

Now if we define our coordinate axes so that the plane of the planetary orbit is in the equatorial plane, then

$$\theta = \frac{\pi}{2}, \qquad \dot{\theta} = 0, \qquad \sin\theta = 1, \qquad \cos\theta = 0.$$

Under these conditions, the components of the metric tensor for this coordinate system become

$$g_{ab} = \begin{bmatrix} \dfrac{1}{\left(1 - \dfrac{2GM}{c^2 r}\right)} & 0 & 0 \\ 0 & r^2 & 0 \\ 0 & 0 & r^2 \end{bmatrix} \qquad (44)$$

When (44) is substituted into (43), two equations result:

$$\frac{1}{\Gamma}\frac{d}{dt}(\Gamma g_{11}\dot{r}) = \frac{1}{2}\frac{\partial g_{11}}{\partial r}\dot{r}^2 + r\dot{\phi}^2 - \frac{GM}{r^2} \qquad (45)$$

$$\frac{d}{dt}(\Gamma r^2\dot{\phi}) = 0. \qquad (46)$$

Eq. (46) expresses the conservation of angular momentum:

$$\Gamma r^2\dot{\phi} = \text{constant} = \frac{l\Gamma}{m_0 g_{11}},$$

where the measured angular momentum l is not a strict constant because the denotation of a "radius vector" used in the definition has an unambiguous meaning only in a flat space. If we substitute for $\dot{\phi}$ in the first equation, drop terms of higher order in v/c, and simplify, we get

$$g_{11}\ddot{r} - \frac{l^2}{m_0^2 r^3 g_{11}^2} + \frac{GM}{r^2} = 0. \qquad (47)$$

Our problem now is to solve this equation, which is usually done by successive approximations. We shall find the task much easier if we use the mathematical shortcut of letting $r = 1/\sigma$ and calculating σ as a function of ϕ rather than t:

$$\frac{dr}{d\sigma} = -\frac{1}{\sigma^2} \qquad \frac{d\phi}{dt} = \frac{l}{m_0 r^2 g_{11}} = \frac{l\sigma^2}{m_0 g_{11}}$$

$$\ddot{r} = \frac{d}{dt}\left(\frac{dr}{dt}\right) = \frac{d\phi}{dt}\frac{d}{d\phi}\left(\frac{d\phi}{dt}\frac{dr}{d\sigma}\frac{d\sigma}{d\phi}\right)$$

$$= -\frac{l^2\sigma^2}{m_0^2 g_{11}^2}\frac{d^2\sigma}{d\phi^2}. \qquad (48)$$

Substituting and rearranging, we get

$$\left(1 - \frac{2GM}{c^2}\sigma\right)\frac{d^2\sigma}{d\phi^2} + \left(1 - \frac{2GM}{c^2}\sigma\right)^2\sigma - \frac{GMm_0^2}{l^2} = 0. \qquad (49)$$

The zeroth approximation neglects all but the two largest terms and leaves us with

$$\sigma_0 = \frac{GMm_0^2}{l^2} = \frac{1}{r_0},$$

which is the equation for a circular orbit of radius r_0.

The first, or Newtonian approximation, neglects the

changes due to the metric tensor:

$$\frac{d^2\sigma}{d\phi^2} + \sigma - \sigma_0 = 0.$$

This equation has the solution

$$\sigma_1 = \sigma_0(1 + e \cos \phi),$$

where e is the eccentricity of the elliptic orbit.

We then put $\sigma = \sigma_1 + \sigma_2$ in (49), and after cancelling out equal terms and neglecting small terms, we obtain

$$\frac{d^2\sigma_2}{d\phi^2} + \sigma_2 - \frac{4GM}{c^2}\sigma_0^2 - \frac{6GM}{c^2}\sigma_0^2 e \cos \phi - \frac{2GM}{c^2}\sigma_0^2 e^2 \cos^2 \phi = 0. \quad (50)$$

Because of the $\cos \phi$ term, we have an equation for a driven oscillator which leads to a continually increasing change of σ with ϕ. Retaining only this term, we find the solution

$$\sigma = \sigma_1 + \sigma_2 = \frac{1}{r_0}\left(1 + e \cos \phi + \frac{3GM}{c^2} e\phi \sin \phi\right) \approx \frac{1}{r_0}\left\{1 + e \cos\left[\phi\left(1 - \frac{3GM}{c^2 r_0}\right)\right]\right\}. \quad (51)$$

It can be seen from (51) that after one revolution, the Newtonian orbit will shift by an amount

$$\Delta\phi = 2\pi \frac{3GM}{c^2 r_0} = \frac{6\pi GM}{c^2 a(1 - e^2)}, \quad (52)$$

where a is the length of the major axis.

This result of Einstein's theory cleared up a bothersome problem in celestial mechanics. The orbit of Mercury is well known and the major axis shifts 5599.74 ± 0.41 sec of arc per century. The perturbations introduced by the other planets in the solar system cause most of this shift, but careful calculations over many years gave the result that the maximum shift due to the planetary perturbations should be 5557.18 ± 0.85 sec of arc per century, leaving a discrepancy of 42.56 ± 0.94 sec of arc per century.

Eq. (52) gives us 42.9 sec of arc per century. This close agreement is a very strong argument in favor of Einstein's equations. Other theories of gravitation, when applied to Mercury's orbit, give an incorrect value or even the wrong sign.

Conclusion

At present, efforts are being made in a number of projects to measure gravitational effects. In the most active of these projects, investigators are attempting to measure the red shift in the frequency of light as it leaves the earth's gravitational field. Cranshaw[13] and others at Harwell, England, and Pound and Rebka[14] at Cambridge have already made measurements using the Mössbauer effect. The results agree with the predictions of general relativity. The Hughes Aircraft Company, National Bureau of Standards, and MIT are working on accurate clocks of various types to put in satellites to measure both special and general relativistic effects. Other experiments to test general relativity using space vehicles have been covered by Benedikt.[15] Kerns[16] at Berkeley and H. E. Fiala at the Hughes Aircraft Company have both made proposals to measure the speed of gravitational interaction. Weber, Zipoy, Forward and Sinsky[17,18] at the University of Maryland are working on the problem of the generation and detection of gravitational radiation.

It is hoped that someone with a practical turn of mind will think of an experiment to detect the gravitational equivalent of the magnetic field. This paper was designed to permit a preliminary evaluation of such ideas.

It is interesting to note that a good electromagnetic autotransformer has almost 100 per cent efficiency in transferring the ac motion of the charges in the primary wire to the charges in the second wire. But the high efficiency is a result of the high velocity of interaction between the charges, the low losses in the wires, and the high permeability of iron. A calculation of the efficiency of a "graviprotation" autotransformer would have to take all these practical considerations into account.

Appendix

Reduction of Einstein's Equations to Show the Analogy to Electromagnetism

The justification for the approximate analogies presented in the main body of the paper is presented here. It is assumed that the reader is familiar with tensor formulation, the summation convention, and the elementary procedures necessary for the handling of Einstein's equation. The procedure for linearizing Einstein's equation is included in all texts on general relativity.[19–21] The calculation of the energy-momentum tensor for slowly moving masses may be found in Møller.[21] The identification of the scalar and vector potentials and the three-dimensional metric tensor and

[13] T. E. Cranshaw, *et al.*, "Measurement of the gravitational red shift using the Mössbauer effect," *Phys. Rev. Lett.*, vol. 4, pp. 163–164; February 15, 1960.

[14] R. V. Pound and G. A. Rebka, Jr., "Apparent weight of photons," *Phys. Rev. Lett.*, vol. 4, pp. 337–341; April 1, 1960.

[15] E. T. Benedikt, "Advances in the Astronautical Sciences," Plenum Press, New York, N. Y., vol. 5, pp. 98–115; 1960.

[16] Q. A. Kerns, "Proposed laboratory measurement of the propagation velocity of gravitational interaction," Lawrence Rad. Lab., Univ. of California, Livermore, Tech. Rept. No. UCRL-8438; December, 1958.

[17] J. Weber, "Detection and generation of gravitational waves," *Phys. Rev.*, vol. 117, pp. 306–313; January 1, 1960.

[18] R. L. Forward, *et al.*, "Upper limit for interstellar millicycle gravitational radiation," *Nature*, vol. 189, p. 473; February 11, 1961.

[19] Landau and Lifshitz, *op. cit.*, p. 324 ff.

[20] A. S. Eddington, "The Mathematical Theory of Relativity," Cambridge University Press, New York, N. Y., p. 128 ff.; 1924.

[21] Møller, *op. cit.*, p. 313 ff.

their use in the equation of motion is also presented by Møller.[22]

We start with Einstein's equations:

$$R_{\alpha\beta} - \frac{1}{2} g_{\alpha\beta} R = \frac{8\pi G}{c^4} T_{\alpha\beta}.$$

Our first assumption will be that all velocities are small, so that special relativity can be neglected, and that all gravitational effects are weak. Then the metric tensor can be approximated by

$$g_{\alpha\beta} \approx \delta_{\alpha\beta} + h_{\alpha\beta},$$

where $\delta_{\alpha\beta}$ is the flat space metric and $h_{\alpha\beta}$ are the perturbations introduced by the masses. Using this form of the metric, the Ricci tensor and the curvature scalar can be calculated from the Christhoffel symbols:

$$R_{\alpha\beta} \approx -\tfrac{1}{2} h_{\alpha\beta}{}^{,\gamma}{}_{,\gamma} = -\tfrac{1}{2} \Box\, h_{\alpha\beta}$$

$$R = g^{\alpha\beta} R_{\alpha\beta} \approx -\tfrac{1}{2} \Box\, \delta^{\alpha\beta} h_{\alpha\beta} = -\tfrac{1}{2} \Box\, h, \qquad (53)$$

where in obtaining (53) we chose our coordinate system so that

$$\left[h_\alpha{}^\beta - \tfrac{1}{2}\delta_\alpha{}^\beta h\right]_{,\beta} = 0.$$

If we substitute the Ricci tensor and the curvature scalar into Einstein's equations, we obtain

$$R_{\alpha\beta} - \tfrac{1}{2} g_{\alpha\beta} R = -\tfrac{1}{2} \Box\, h_{\alpha\beta} + \tfrac{1}{4}\delta_{\alpha\beta} \Box\, h = \frac{8\pi G}{c^4} T_{\alpha\beta}. \quad (54)$$

We now define the gravitational potential as

$$\phi_{\alpha\beta} = h_{\alpha\beta} - \tfrac{1}{2}\delta_{\alpha\beta} h;$$

substituting and rearranging, we get

$$\Box\, \phi_{\alpha\beta} = -\frac{16\pi G}{c^4} T_{\alpha\beta}.$$

If we write out the d'Alembertian operator, we have

$$\Delta\phi_{\alpha\beta} - \frac{1}{c^2}\frac{\partial^2}{\partial t^2}\phi_{\alpha\beta} = -\frac{16\pi G}{c^4} T_{\alpha\beta}. \qquad (55)$$

This is the basic equation upon which the analogies are based.

Scalar Potential

In the first approximation, we assume that all quantities are not varying with time and that the masses have low velocities or rotations. Then the time derivative of the gravitational potential is zero and all the components of the energy-momentum tensor are zero except

$$T_{oo} = \mu c^2.$$

Eqs. (55) reduce to

$$\Delta\phi_{oo} = -\frac{16\pi G}{c^2}\mu. \qquad (56)$$

This is essentially Poisson's equation, which has the solution

$$\phi_{oo} = +\frac{4G}{c^2}\int_V \frac{\mu}{r}\, dV.$$

If we define a gravitational capacitivity of the vacuum as

$$\gamma = (4\pi G)^{-1},$$

we get

$$-\frac{c^2\phi_{oo}}{4} = -\frac{1}{4\pi\gamma}\int_V \frac{\mu}{r}\, dV. \qquad (57)$$

By comparing (57) with the scalar potential of an electric charge density

$$\phi = +\frac{1}{4\pi\epsilon}\int_V \frac{\rho}{r}\, dV,$$

we see that we can construct the well-known gravitational analog to the scalar potential:

$$\chi = -\frac{c^2\phi_{oo}}{4} = -\frac{c^2(g_{oo}+1)}{2}.$$

Space Curvature

This first approximation (56) also determines the spatial metric. The existence of the component ϕ_{oo} results in an interval of the form

$$ds^2 = \left(1 - \frac{2\chi}{c^2}\right)(dx^2 + dy^2 + dz^2) - \left(1 + \frac{2}{c^2}\chi\right)c^2dt^2. \qquad (58)$$

Thus the three-dimensional spatial metric will be of the form

$$g_{ab} = \begin{vmatrix} 1 - \frac{2\chi}{c^2} & 0 & 0 \\ 0 & 1 - \frac{2\chi}{c^2} & 0 \\ 0 & 0 & 1 - \frac{2\chi}{c^2} \end{vmatrix}.$$

In higher approximations that will be considered later, the additional terms in the spatial metric will be smaller than $2\chi/c^2$ by the order of $(v/c)^2$, and since we assume velocities much smaller than the speed of light, they will be of little experimental interest.

Vector Potential

In the next higher approximation, we still assume that the potential is not varying with time, but that the masses involved have appreciable velocity or rota-

[22] *Ibid.*, p. 246 ff. and p. 288 ff.

tion. Then the energy-momentum tensor will have the components

$$T_{oo} = \mu c^2 \left(\text{zero order in } \frac{v}{c}\right)$$

and

$$T_{ao} = -\mu c^2 \left(\frac{v_a}{c}\right) \left(\text{first order in } \frac{v}{c}\right).$$

We then have four equations remaining: one gives us the scalar potential obtained previously, and the other three are

$$\Delta\phi_{ao} = +\frac{16\pi G}{c^3}\mu v_a.$$

These equations have the solution

$$\phi_{ao} = -\frac{4G}{c^3}\int_V \frac{\mu v_a}{r}\,dV.$$

If we define a gravitational permeability of space by

$$\eta = \frac{16\pi G}{c^2},$$

then we can substitute and rearrange to get

$$c\phi_{ao} = -\frac{\eta}{4\pi}\int_V \frac{\mu v_a}{r}\,dV.$$

Thus we can identify a mass density flow $\boldsymbol{p} = \mu \boldsymbol{v}$ and a gravitational equivalent of the vector potential whose components are the three components of

$$K_a = c\phi_{ao} = cg_{ao},$$

and thereby arrive at the isomorphism of the equations

$$\boldsymbol{K} = -\frac{\eta}{4\pi}\int_V \frac{\boldsymbol{p}}{r}\,dV \qquad \boldsymbol{A} = \frac{\mu}{4\pi}\int_V \frac{\boldsymbol{j}}{r}\,dV. \qquad (59)$$

Gravitational Radiation

Let us return to the basic equation (55):

$$\Delta\phi_{\alpha\beta} - \frac{1}{c^2}\frac{\partial^2}{\partial t^2}\phi_{\alpha\beta} = -\frac{16\pi G}{c^4}T_{\alpha\beta}. \qquad (60)$$

As it stands, (60) is a wave equation for the gravitational potential $\phi_{\alpha\beta}$. The velocity of propagation is the same as the velocity of light. It is this equation, which results from the linearization of Einstein's equation, that gives credence to the statement that gravitational radiation exists. The solution of the wave equation and the calculation of the radiated power are straightforward and are carried out in Landau and Lifshitz.[19]

Equation of Motion

In the four-dimensional equation of motion (61), the gravitational effects are entirely in the metric tensor. The only forces explicitly stated are nongravitational forces. This four-dimensional equation of motion can be broken down and arranged so that it is a three-dimensional curvilinear spatial equation of motion. The gravitational effects resulting from the temporal components of the metric tensor are represented as forces due to a gravitational scalar and a gravitational vector potential. The spatial components of the metric tensor are used as the three-dimensional metric tensor.

The general equation of motion for a particle with only gravitational forces acting is given by Møller[23] as

$$\frac{dP_\alpha}{d\tau} - \frac{1}{2}\frac{\partial g_{\beta\gamma}}{\partial x^\alpha}U^\beta P^\gamma = 0, \qquad (61)$$

where

$$P_\alpha = m_0 U_\alpha = m_0 g_{\alpha\beta}U^\beta = m_0 g_{\alpha\beta}\Gamma\frac{dx^\beta}{dt}$$

and

$$\Gamma = \frac{dt}{d\tau} = \left\{\left[\left(1+\frac{2\chi}{c^2}\right)^{1/2} - \frac{K_a v^a}{c^2}\right]^2 - \frac{v^2}{c^2}\right\}^{-1/2}$$

The $\alpha = 0$ equation gives us the conservation of mass-energy and the other three equations are

$$\begin{aligned}\Gamma\frac{d}{dt}\left(m_0\Gamma g_{ab}\frac{dx^b}{dt}\right) &+ \Gamma\frac{\partial}{\partial t}\left(m_0\Gamma g_{ao}\frac{dx^o}{dt}\right)\\ &+ \Gamma\frac{dx^b}{dt}\frac{\partial}{\partial x^b}\left(m_0\Gamma g_{ao}\frac{dx^o}{dt}\right)\\ &- \frac{1}{2}\frac{\partial g_{oo}}{\partial x^a}m_0\Gamma^2\left(\frac{dx^o}{dt}\right)^2 - \frac{\partial g_{bo}}{\partial x^a}m_0\Gamma^2\frac{dx^o}{dt}\frac{dx^b}{dt}\\ &- \frac{1}{2}\frac{\partial g_{bc}}{\partial x^a}m_0\Gamma^2\frac{dx^b}{dt}\frac{dx^c}{dt} = 0.\end{aligned}$$

Dividing through by $m_0\Gamma^2$, letting $dx^o/dt = c$, and neglecting higher-order terms, we get

$$\begin{aligned}\frac{1}{\Gamma}\frac{d}{dt}\left(\Gamma g_{ab}\frac{dx^b}{dt}\right) &= \frac{1}{2}\frac{\partial g_{bc}}{\partial x^a}\frac{dx^c}{dt}\frac{dx^b}{dt} + \frac{c^2}{2}\frac{\partial g_{oo}}{\partial x^a}\\ &\quad - c\frac{\partial g_{ao}}{\partial t} + c\left(\frac{\partial g_{bo}}{\partial x^a} - \frac{\partial g_{ao}}{\partial x^b}\right)\frac{dx^b}{dt}.\end{aligned}$$

We then use our definition of the gravitational scalar and vector potential

$$K_a = cg_{oa}$$

$$v^a = \frac{dx^a}{dt}$$

$$\chi = -\frac{c^2}{2}(g_{oo}+1)$$

$$g_{oo} = -1 - \frac{2\chi}{c^2}$$

[23] Møller, *op. cit.*, p. 290.

to get

$$\frac{1}{\Gamma}\frac{d}{dt}\left(\Gamma g_{ab}\frac{dx^b}{dt}\right) = \frac{1}{2}\frac{\partial g_{bc}}{\partial x^a}v^b v^c - \frac{\partial \chi}{\partial x^a} - \frac{\partial K_a}{dt} + v^b\left[\frac{\partial K_b}{\partial x^a} - \frac{\partial K_a}{\partial x^b}\right]. \tag{62}$$

The left-hand side of (62) is the acceleration of the particle in a curvilinear coordinate system. The first term on the right gives the fictional forces due to the choice of the coordinate system, the most familiar examples being the coriolis force and the centrifugal force. The second term is one component of $\nabla\chi$, the gravitational static attraction; the third term is one component of $\partial \boldsymbol{K}/\partial t$, the gravitational induction effect; and the fourth term is one component of $\boldsymbol{v}\times(\nabla\times\boldsymbol{K})$, the gravitational equivalent of the Lorentz force.

A Matched Amplifier Using Two Cascaded Esaki Diodes*

DONALD R. HAMANN†

Summary—**The purpose of this paper is to introduce a new type of circuit for matched amplification using negative resistance devices. This circuit consists of a quarter-wave transmission line section whose input and output are paralleled by negative conductances. The characteristics of such an amplifier are discussed, and an expression for its noise figure is derived. The development of a 30-Mc amplifier using two Esaki diodes is described. Experimental results are presented, including curves of the characteristics as a function of frequency. A gain of 8.9 db was measured with a noise figure of 4.3 db.**

Introduction

AMPLIFICATION by a single negative resistance element has several disadvantages. First, such an amplifier necessarily presents a mismatch to the source. Second, amplification and reflection of noise power radiated by the load tends to contribute to its noise figure. The noise figure of such an amplifier is given by (1).

$$F = 1 + \frac{\overline{|i|^2}}{4kT_SG_S} + \frac{G_LT_L}{G_ST_S}. \tag{1}$$

In this expression, $\overline{|i|^2}$ is the equivalent noise current squared per cycle produced by the negative resistance device, G_S and G_L are the source and load conductances, and T_S and T_L are the source and load temperatures. It may be seen that in order to achieve a low noise figure, the factor G_LT_L must approach zero, regardless of the device used. If G_L becomes small, however, it is necessary to adjust the source and device conductances extremely close to the point of oscillation in order to achieve any power gain.[1]

These effects may be eliminated by the use of nonreciprocal circuit elements, or a hybrid with two negative resistance elements.[2,3] At lower frequencies, nonreciprocal passive elements are difficult to produce, and hybrids do not present the wide frequency range resistive load necessary to stabilize Esaki diodes. This paper describes another method of obtaining amplification using two negative resistance elements. The amplifier described is matched to both the source and load, and achieves a low noise figure with equal source and load conductances. Experimental work was done at 30 Mc, permitting the use of lumped circuits. Although Esaki diodes were used, the circuit techniques should be equally applicable to parametric and other negative resistance devices.

Theory of Operation

The basic circuit which was used consists of two conductances G and a $\frac{1}{4}$ wave transmission line section of impedance Z, as shown in Fig. 1. A 1-ohm source and load are assumed to simplify computation, so that the factors G and Z are normalized variables in all equations. The circuit may be easily analyzed using the fourpole matrix notation

$$\begin{Bmatrix} E_{\text{in}} \\ I_{\text{in}} \end{Bmatrix} = \begin{pmatrix} A & B \\ C & D \end{pmatrix}\begin{Bmatrix} E_{\text{out}} \\ I_{\text{out}} \end{Bmatrix}. \tag{2}$$

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† Mass. Inst. Tech., Cambridge, Mass.
[1] K. K. N. Chang, "Low noise tunnel-diode amplifier," Proc. IRE, vol. 47, pp. 1268–1269; July, 1959.

[2] M. E. Hines, "High-frequency negative-resistance circuit principles for Esaki diode applications," *Bell Sys. Tech. J.*, vol. 39, pp. 485–488; May, 1960.

[3] L. U. Kibler, "Directional bridge parametric amplifier," Proc. IRE, vol. 47, pp. 583–584; April, 1959.

The matrix representing the network shown in Fig. 1 is

$$\begin{pmatrix}1 & 0\\ G & 1\end{pmatrix}\begin{pmatrix}0 & jZ\\ j/Z & 0\end{pmatrix}\begin{pmatrix}1 & 0\\ G & 1\end{pmatrix}=\begin{pmatrix}jGZ & jZ\\ j[1/Z+G^2Z] & jGZ\end{pmatrix}. \quad (3)$$

A normalized fourpole is matched if $A=D$ and $B=C$, a requirement which is satisfied when

$$Z^2=\frac{1}{1-G^2}. \quad (4)$$

The insertion power gain of a fourpole is

$$g=\frac{4}{|A+B+C+D|^2}. \quad (5)$$

If this expression is evaluated for the network under discussion, and the match requirement given by (4) is imposed, it is found that

$$g=\left|\frac{1-G}{1+G}\right|. \quad (6)$$

This expression shows that g becomes greater than unity when G becomes negative, and becomes large when G approaches -1. For each value of g, there are two solutions for G. If the solution for which $G>-1$ is chosen, the total conductance connected to the reactive elements of the amplifier is always positive, as is shown in Fig. 2. Thus, although the network may only display matched gain in the neighborhood of the design frequency, it will be stable at all frequencies.

Matched amplification can be achieved by a number of circuits equivalent to Fig. 1. The equivalent shown in Fig. 3 consisting of an ideal transformer, a series resistance GZ^2, and a shunt conductance G, is of particular interest. It shows that all reactive elements, and hence all frequency dependence, can in principle be removed from such circuits. Stabilization difficulties prevented extensive experimental work with this circuit.

Since these amplifiers are matched, their noise figure, F, may be calculated quite simply using the familiar Friis formula giving

$$F=1+\frac{\overline{|i_1|^2}}{4kT_S}+\frac{1}{g}\left(\frac{\overline{|i_2|^2}}{4kT_S}+\frac{T_L}{T_S}\right). \quad (7)$$

In this expression, i_1 and i_2 are the noise sources associated with the input and output negative conductance devices respectively. It is seen that contributions to the system noise figure from the output device, and from the load noise, become small as the gain increases.

Experimental Results

An experimental amplifier was built based upon the circuit in Fig. 1 operating at 30 Mc with a 50-ohm source and load. A more detailed schematic is shown in Fig. 4. L_L and C_L formed a lumped $\frac{1}{4}$ wave line. Since available diodes had a maximum negative conductance of -0.0086 mho, transformers T_1 and T_2 were necessary to obtain the desired negative conductance of approximately -0.02 mho for high gain. The final adjustment of this parameter was made by varying bias voltages E_1 and E_2. The networks R_s and C_s provided a load for the diodes at higher frequencies, at which the transformers no longer coupled the diodes to the remainder of the circuit.

The amplifier was adjusted for a gain of 8.9 db with little difficulty. Operating with this gain, it could be matched well enough to provide at least a 30-db return loss at either terminal pair with the other exactly terminated. At the design frequency the amplifier should

Fig. 1—Basic circuit of the matched amplifier.

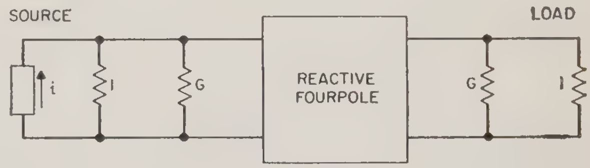

Fig. 2—Amplifier showing inherent stability under loading.

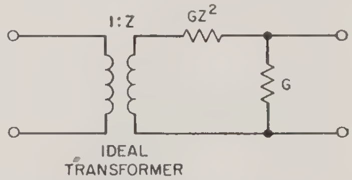

Fig. 3—Frequency insensitive equivalent of matched amplifier.

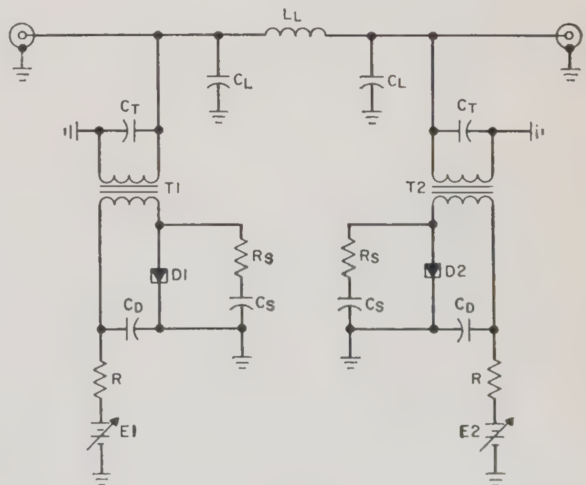

Fig. 4—Experimental amplifier.

be able to tolerate a wide range of mismatches in both source and load without oscillation. It is required that the sum of the source and load return losses less twice the gain be positive for stable operation.

The particular form of the amplifier which was constructed showed one undesirable characteristic. At frequencies above and below 30 Mc it exhibited peaks of transmission and return gain. Mismatch in the source and load at these frequencies made it impossible to obtain a higher 30-Mc gain without oscillation. Both these characteristics were predicted by the analysis of a somewhat simplified model of the circuit shown in Fig. 4. This model included L_L, C_L, C_T, T_1, T_2, D_1 and D_2, and assumed the diodes to be ideal negative conductances and the transformers to have unity coupling coefficients. The theoretical and measured values of the gain and return gain as functions of frequency are shown in Figs. 5 and 6. It may be observed that a fairly wide flat region of gain exists around 30 Mc, and that a ten per cent bandwidth is obtained for a return gain of less than -10 db.

By assuming that the diodes produce shot noise, and evaluating $\overline{|i|^2}$ as in (8), the noise figure can be calculated from (7).

$$\overline{|i|^2} = 2eI \frac{G}{G_d}. \tag{8}$$

In (8), I is the bias current, G_d the actual diode conductance, and G the network transformation of G_d appearing in (4). The factor G/G_d accounts for both physical transforming networks and normalization to unit impedance. Operating under the above conditions, a zero load temperature noise figure of 3.2 db was calculated, and one of 4.3 db was measured. It is believed that the discrepancy was due to an oversimplified model of the Esaki diode noise source. No attempt was made to find the optimum low noise operating point on the diode characteristics.

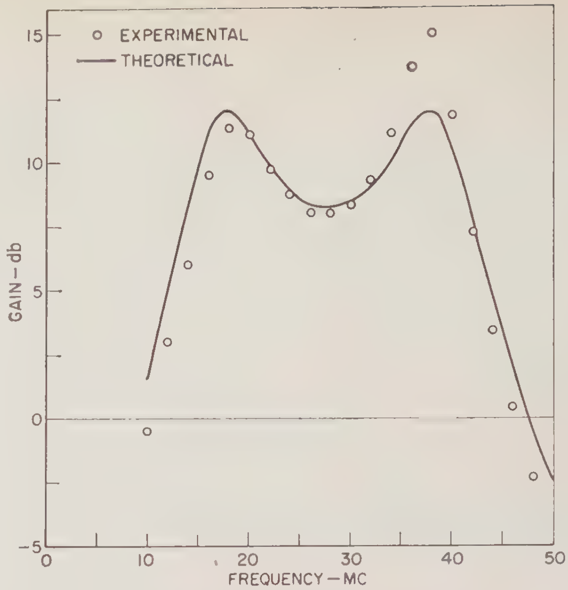

Fig. 5—Transmission gain of circuit of Fig. 4 as a function of frequency.

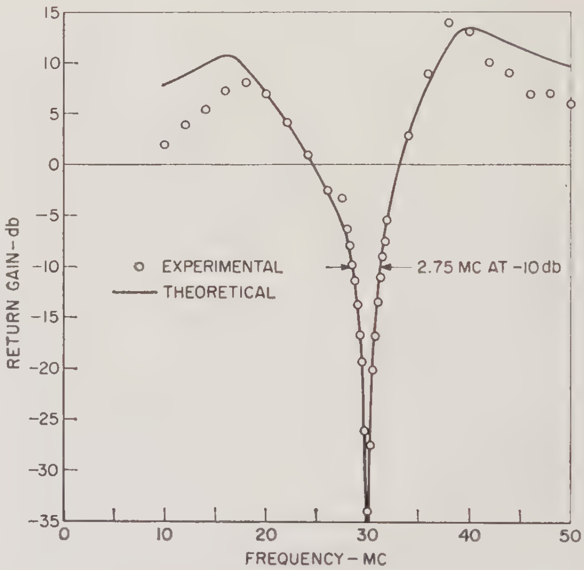

Fig. 6—Return gain as a function of frequency, measured at one terminal pair with the other terminated.

Conclusions

In the high gain limit, (7) for the two-diode amplifier and (1) for the single-diode amplifier lead to an identical noise figure formulation

$$F = 1 + \frac{eI}{2kT_S} \frac{G}{G_d}. \tag{9}$$

This result is true, although the diode pair system works into a matched load, whereas the single diode requires an open circuit termination. The limitation of performance by the I/G_d ratio remains basic and does not mitigate the noise requirements of the diode.

The usefulness of the form of the circuit shown in Fig. 1 has been established. Its undesirable regenerative characteristics at frequencies removed from the design frequency would not be present if a narrow band negative resistance, such as a parametric amplifier, were used. Alternately, the inherent stability of the circuit suggests that bandwidth may be improved by changing the form of the reactive fourpole. The form of the network shown in Fig. 3 promises even greater bandwidth, but may be subject to unavoidable instabilities introduced by parasitic reactances. Further work in these directions should produce a simple and useful low noise amplifier without the undesirable characteristics of present negative-resistance amplifiers.

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Network Synthesis with Negative Resistors*

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Summary—**The development of new solid-state active elements, such as variable-capacitor diodes and tunnel diodes, has stimulated the network theorist to consider the negative resistor as an additional basic circuit element to be included in problems of linear network analysis and synthesis. In this paper it is first shown that if the negative resistor is added to the usual set of lumped passive building blocks, then it is possible to represent as a network *any* linear relation between *n*-port voltages and currents prescribed in terms of real, rational functions of a complex-frequency variable. This leads to the synthesis of some novel pathologic circuits which have neither immittance nor scattering representations, such as a one-port, which is simultaneously an open circuit and a short circuit ($v=i=0$, the "nullator"), and the linear network in which voltages and currents at the ports are completely arbitrary (the "norator," the unique, linear nonreciprocal, one-port). These elements are shown to be basic linear circuit building blocks. The second part of the paper considers the synthesis in the frequency domain of a real, rational $n\times n$ immitance matrix in which pole locations and pole multiplicities are completely arbitrary. It is shown that such a matrix can always be realized with lossless elements and at most *n* positive and *n* negative resistors.**

Summary of Negative-Resistance Network Techniques

THIS PAPER treats the properties and synthesis of linear networks which contain one or more of six different kinds of idealized lumped network elements. The elements are the positive inductor, positive capacitor, positive resistor, ideal transformer, ideal gyrator and negative resistor. The inclusion of the last two elements means that the domain of discussion encompasses linear nonreciprocal and active lumped networks.

Two closely related subjects are considered. The first is a general method for describing network terminal properties which is valid even when the usual immittance and scattering formalisms do not exist. This more general formalism is necessary in the active domain because pathologic circuits are encountered which are not met with in the more familiar reaches of passive network theory. A final result of the first section on representation is a technique for appropriately operating on a network description so that pathologies are removed and an immittance or scattering matrix is obtained even if neither of these existed initially. This method, for example, can lead to equivalent circuits for the degenerate network which simultaneously exhibits the properties of a short circuit and an open circuit at its input port. The technique is applicable whether the network parameters are constants or rational functions of frequency. In effect, this first portion of the paper shows that *any* prescribed linear relations between voltages and currents of an *n* port which is given in the complex-frequency domain in terms of rational functions with real coefficients can be constructed as a network graph with passive lumped elements and negative resistors.

The second section of the paper discusses the synthesis of networks when an immittance (or scattering) description has been made available, say, by the methods of the first section. A discussion is given of one-port and *n*-port network synthesis techniques utilizing a minimum number of positive and negative resistors. It is shown that an *n* port requires no more than *n* positive and *n* negative resistors for its synthesis.

Representation of Negative-Resistance Networks—Pathologic Circuits, Nullators, Norators

The network formalism chosen here for describing the linear volt-ampere relations at accessible ports applies to all networks which have a graph.[1] The formalism is that introduced by Belevitch [1]. In this description, the linear relations between voltages and currents at the ports of an *n*-port network are depicted as

$$AV = BI, \tag{1a}$$

or

$$[A, -B]\begin{bmatrix} V \\ I \end{bmatrix} = 0. \tag{1b}$$

Here A and B are $n\times n$ square matrices, and V and I are single-column n-element matrices of voltage and current variables at the n accessible ports. We first indicate how (1b) may define immittance or scattering representations.

Clearly an immittance representation can be obtained from (1) if A or B is nonsingular. Then

$$Z = A^{-1}B, \quad (A \text{ nonsingular}) \tag{1c}$$

$$Y = B^{-1}A. \quad (B \text{ nonsingular}) \tag{1d}$$

If the ports of the network are series augmented (*i.e.*, elements added in series with the ports) with resistors $r_1, r_2, \cdots, r_n$, the voltage and current vectors V', I' at the new terminals beyond the series-connected re-

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[1] Any linear relationship between $2n$ variables, the V and I at n prescribed ports, whose coefficients are functions of a complex-frequency variable, can be expressed in the form of (1b). This equation is equivalent to the existence of a linear vector space of order $2n$ and dimensionality r, with $0\leq r\leq 2n$. Thus any linear n port with a graph must satisfy (1b). In the general case, A and B are $m\times n$, $m\neq n$.

sistors are related to the original variables V, I by

$$I = I'$$
$$V = V' - RI',$$

where R is a diagonal matrix indicated by diag $(r_1, r_2, \cdots, r_n)$. Substituting in (1),

$$AV' = (AR + B)I'. \tag{2}$$

Then if R can be found so that $(AR+B)$ is nonsingular, the admittance of the augmented network is

$$Y_A = (AR + B)^{-1}A. \tag{3a}$$

Similarly, shunt augmenting with parallel conductances $G = \text{diag}(g_1, g_2, \cdots)$ can result in

$$Z_A = (A + BG)^{-1}B, \tag{3b}$$

if $(A+BG)$ is nonsingular. If (3) applies, a normalized scattering matrix can be obtained:

$$S = E_n - 2R^{1/2}Y_AR^{1/2}, \tag{4a}$$

$$S = E_n - 2G^{1/2}Z_AG^{1/2}, \tag{4b}$$

where E_n is an $n \times n$ identity matrix, $R^{1/2} = \text{diag}\,[\sqrt{r_1}, \sqrt{r_2}, \cdots, \sqrt{r_n}]$, and similarly for $G^{1/2}$.

The discussion which follows is intended to show that the representation of (1b), when it contains real, rational elements of a complex-frequency variable, always leads to a network graph containing the six basic elements mentioned in the opening paragraph. It should be pointed out that in this portion of the paper no attempt is made to demonstrate a minimal synthesis; we merely show that a network representation always exists. The first theorem presents the conditions under which it is possible to obtain a synthesis by merely adding resistors to the ports of the unknown n port to produce a Z or Y matrix which is directly amenable to network representation. Following this, a more general procedure is obtained in cases where series-resistor augmentation cannot lead to a synthesis. Thus Theorem 2 considers the case where the A and B matrices are square and $[A, B]$ is of rank n. Finally, Theorem 3 shows that any $m \times 2n$ $[A, B]$ matrix of rational fractions with real coefficients may be synthesized as an n port with passive lumped elements and negative resistors no matter whether $[A, B]$ is of rank n or not, and whether or not A and B are square. In effect, this says that in the complex-frequency domain, any linear, rational description with real coefficients of an n port can be realized with the six basic lumped elements. It should be emphasized that this general theorem leads to the synthesis of extremely pathologic and novel networks. One of these is the one port which is simultaneously a short circuit and an open circuit, $(v = i = 0)$, as well as the one port in which v and i are completely arbitrary. Though space here does not permit a philosophic discussion of such "odd-ball" networks, one cannot deny their existence, provided the six idealized elements are postulated.

We now proceed to the case where immittance or scattering formalisms need not exist.

If the normalization numbers r_k or g_k of R or G in (4) are unity, S will exist if $A+B$ is nonsingular. In general, this will always be true *if the network is passive*, but is not necessarily true if the network is active. In the case of active circuits, A, B, $A+B$ may be simultaneously singular, and indeed it may not be possible to find a set of numbers $r_1, r_2, \cdots, r_n$ so that $(AR+B)$ is nonsingular. Thus a possible degeneracy which may occur with negative-resistance circuits is a network description for which neither Z, Y nor S exists. As an example of such a situation we have

$$AV = BI = \begin{bmatrix} 1 & 0 \\ 0 & 0 \end{bmatrix}\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ 1 & 0 \end{bmatrix}\begin{bmatrix} i_1 \\ i_2 \end{bmatrix}. \tag{5}$$

If a set of numbers defining R can be found such that $(AR+B)$ is nonsingular ($A+BG$ will not be considered separately here since the two cases are exactly dual), then Y_A may be found according to (3a), and the network N corresponding to $[A, -B]$ can be constructed by synthesizing Y_A and then series augmenting the resultant structure by $-R$ to get N. Further, if all the r_k are positive, an S with nonidentical positive-port normalization numbers may be found and the network N synthesized from this S [2]. Accordingly, the conditions under which R may be determined so that a nonsingular matrix $(AR+B)$ may be constructed will now be investigated. It will be apparent that if such an R exists, all its elements may always be chosen positive so that synthesis from Y_A or S can follow from an identical set of restrictions on A and B.

Consider the following equation:

$$AR + B = [A, B]\begin{bmatrix} R \\ E_n \end{bmatrix}. \tag{6}$$

It is clear since

$$\begin{bmatrix} R \\ E_n \end{bmatrix}$$

is of rank n, that by the rule of rank if $(AR+B)$ is to be nonsingular, it is, at the least, necessary for $[A, B]$ to be of rank n. This condition, however, is not sufficient, and we must establish the requirements under which an n column minor of $(AR+B)$ will not vanish. To do this we use the Binet-Cauchy [3] theorem for the determinant of the product of two rectangular matrices. Here, they are $n \times 2n$ and $2n \times n$, respectively. This product is equal to the sum of products of minors, formed from each. Each term of this sum is the product of an n-column minor of the first matrix, $[A, B]$, by the corresponding n-rowed minor in

$$\begin{bmatrix} R \\ E \end{bmatrix}.$$

Clearly, at least one of these products must be shown to be non-zero for $AR+B$ to be nonsingular. If more than one term is non-zero, the positive elements r_k can be adjusted so that the sum of two or more non-zero terms does not vanish. In order that at least one of the terms does not vanish, a non-zero n-rowed minor of

$$\begin{bmatrix} R \\ E \end{bmatrix}$$

must be paired with a similar non-zero n-column minor of $[A, B]$. But all n-rowed minors of

$$\begin{bmatrix} R \\ E \end{bmatrix}$$

formed of rows of R and noncorrespondingly-ordered rows of E are non-zero, all other minors are zero. Thus if $n=3$, the following minors do not vanish: rows 1, 2, 3, of R; rows 1, 2 of R, 3 of E; row 1 of R; 2, 3 of E; row 1, 3 of R, 2 of E; etc. When minors are chosen in this fashion, they will be termed "complementary n-rowed minors" of

$$\begin{bmatrix} R \\ E \end{bmatrix}$$

and are formed of "complementary rows of R and E." One can then phrase the rule that the determinant of $(AR+B)$ be non-zero: a complementary n-column minor of $[A, B]$ (*i.e.*, formed of complementary columns of A and B) be non-zero. (A matrix with this property has been termed "special" [4].) If this is true, this minor will automatically be paired with a complementary n-rowed, hence non-zero, minor of

$$\begin{bmatrix} R \\ E \end{bmatrix}.$$

Thus we have Theorem 1.

Theorem 1: The necessary and sufficient condition that:

1) An immittance matrix description be constructed from an $n\times 2n$ $[A, B]$ matrix of real, rational functions[2] by adding positive resistors r_k to the n ports of the network; or

2) A rational, real scattering matrix with positive-port normalization numbers r_k be formed from $[A, B]$;

$[A, B]$ contain a nonvanishing n-column minor formed of complementary columns from A and B.

[2] The process of constructing the immittance or scattering matrix from $[A, B]$ as described here involves operating on A and B with *matrices* of constants independent of the functional form of the elements of $[A, B]$. These may therefore be, for example, constants or rational functions.

As an example consider the defining equations

$$V_1 + V_2 = 0,$$

$$I_1 + I_2 = 0,$$

or

$$\begin{bmatrix} 1 & 1 & 0 & 0 \\ 0 & 0 & 1 & 1 \end{bmatrix}\begin{bmatrix} V \\ I \end{bmatrix} = 0.$$

Note that A, B, and $A+B$ are singular. (This is not a passive network.) However, the complementary minor of $[A, B]$ formed from the first column of A and the second column of B,

$$\begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix},$$

does not vanish. Hence, an appropriate R can be found. Thus:

$$(AR + B) = \begin{bmatrix} 1 & 1 \\ 0 & 0 \end{bmatrix}\begin{bmatrix} r_1 & 0 \\ 0 & r_2 \end{bmatrix} + \begin{bmatrix} 0 & 0 \\ 1 & 1 \end{bmatrix} = \begin{bmatrix} r_1 & r_2 \\ 1 & 0 \end{bmatrix},$$

and any $r_1 \neq r_2$, both positive, will produce a nonsingular $(AR+B)$. The realization of this circuit which requires positive and negative resistors and a gyrator and is obtained from the admittance given by (3a) is an equivalent for a negative impedance converter and may be carried out as described in Reference [5].

The next more general case to consider in transforming the $[A, B]$ network description so as to obtain an immittance or scattering representation is that in which $[A, B]$ has no non-zero complementary n-column minors but is still of rank n. An example of this was give in (5). Consider the following:

$$[A, B]\begin{bmatrix} V \\ I \end{bmatrix} = [A, B][PP^{-1}]\begin{bmatrix} V \\ I \end{bmatrix}$$

$$= [A' \ B']\begin{bmatrix} V' \\ I' \end{bmatrix}. \tag{7}$$

P is a nonsingular $2n\times 2n$ matrix which performs the elementary operation of interchanging columns of $[A, B]$, and as such, is its own inverse. Thus also:

$$P^{-1} = P \tag{8}$$

$$[A', B'] = [A, B]P \tag{9a}$$

$$\begin{bmatrix} V' \\ I' \end{bmatrix} = P\begin{bmatrix} V \\ I \end{bmatrix}$$

or

$$\begin{bmatrix} V \\ I \end{bmatrix} = P\begin{bmatrix} V' \\ I' \end{bmatrix}. \tag{9b}$$

We will now show that if $[A, B]$ is of rank n, P may always be chosen so that A' in $[A', B']$ is nonsingular and thus an impedance matrix defined by (1c) exists.

Then $[A', B']$ may be synthesized from this impedance description. Finally, a synthesis for $[A, B]$ may be obtained if the operation described by (9b) is interpreted. This will be done by constructing a $2n$-port network corresponding to P, n ports of which are connected to the $[A', B']$ network with V, I defined at the remaining accessible n ports. Thus the terminal quantities V', I' are transformed to V, I.

The first part of this demonstration is quite simple. $[A, B]$ is of rank n, hence it has n independent columns. The column permuting matrix P is then chosen so that the n columns of $[A, B]$ which form a nonvanishing determinant are rearranged in position to occur as the first n columns of $[A', B']$, *i.e.*, form A'. Thus A' is of rank n, and by (1c), the network corresponding to $[A', B']$ has an impedance matrix $Z' = (A')^{-1}B$. The $[A', B']$ network may therefore be synthesized from Z' if $[A, B]$ contains rational elements [2].

It remains to be shown that the transformation effected by P has a physical representation. The matrix P in (9b) is an impedance transfer matrix of constants between the output ports

$$\begin{bmatrix} V' \\ I' \end{bmatrix}$$

and the input ports

$$\begin{bmatrix} V \\ I \end{bmatrix}$$

as shown in Fig. 1, and it is demonstrated in Appendix I that the $2n$-port network associated with P in that figure has an $[A, B]$ representation which satisfies Theorem 1.

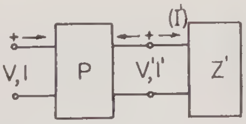

Fig. 1—Transformation of an impedance matrix.

Hence, P is realizable and thus we have

Theorem 2:

> A sufficient condition that an $n \times 2n$ $[A, B]$ matrix of rational functions with real coefficients be represented by an n-port lumped network (*i.e.*, containing L, C, $\pm R$, gyrators and ideal transformers) is that $[A, B]$ contain real, rational functions and be of normal rank n.

As in Theorem 1, operations on A and B were with matrices of constants independent of the functional form of the elements of $[A, B]$; hence, even if $[A,B]$ contains real, rational functions, the transformation process may still be carried out as stated in Theorem 2.

As an example of the technique just described, consider (5). If this is expanded, we obtain as defining equations a two port,

$$v_1 = 0, \qquad i_1 = 0,$$
$$v_2,\ i_2, \text{ arbitrary.} \tag{10}$$

The anomalous two port thus described performs simultaneously as a short circuit and open circuit at port one. Oono has discussed this peculiar circuit [6] and has given an extensive treatment of a minimal realization of two ports described by an $[A, B]$ matrix of constants of which the degenerate two port of (10) is a special case. Belevitch has presented one realization of (10) [7]. The discussion in this paper differs from Oono's in that it is simpler to apply for n ports and is not restricted to an $[A, B]$ matrix of constants, but it is not minimal in the number of elements. Belevitch does not present an organized synthesis procedure.

Returning to our example, the $[A, B]$ matrix is

$$[A, B] = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix}.$$

If columns two and three are interchanged, the result is an $[A', B']$ matrix which has an impedance description. Thus from (9a),

$$[A', B'] = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix} = [A, B] \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix}$$
$$= [A, B]P;$$

and by (1c),

$$Z' = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ 0 & 0 \end{bmatrix}.$$

This is a two port consisting of two isolated short circuits. The $[A, B]$ matrix corresponding to P as given in the above example may be written down from (47b) [Appendix I].

$$[A_P, B_P] = \left[\begin{array}{cccc:cccc} 1 & \dot{0} & -\dot{1} & 0 & \dot{0} & 0 & 0 & \dot{0} \\ 0 & 1 & 0 & 0 & 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 & 0 & 1 & 0 & -1 \end{array}\right]$$

Columns 2, 3, 5, and 8 form a nonvanishing complementary minor. We may augment the network corresponding to $[A, B]_P$ with resistors to obtain an admittance representation, and then introduce the negative of these resistors in series with the ports to get a $2n$-port realization for P. In this case, it is convenient (though not necessary) to choose some negative values for the augmenting resistors. Thus for R as used in (2), choose

$$R = \operatorname{diag}(-1, 1, -1, -1).$$

The resulting augmented Y_A matrix is then, by (3a),

$$Y_A = (A_P R + B_P)^{-1} A_P$$
$$= \frac{1}{2}\begin{bmatrix} 1 & 1 & -1 & -1 \\ -1 & -1 & 1 & -1 \\ -1 & 1 & 1 & -1 \\ 1 & 1 & -1 & 1 \end{bmatrix}.$$

This network is then synthesized [5], the resistors obtained from $-R$ are placed in series with the four ports to give the representation of P, and the two short circuits corresponding to Z' placed across ports 3 and 4 to realize the $[A, B]$ matrix originally specified at ports 1 and 2. The resulting network as determined by the process just described is shown on Fig. 2.

The existence of a network which is simultaneously an open and a short circuit at a given port is demonstrated by the actual structure shown in Fig. 2. However it must be pointed out that while the synthesis technique gives a realization, the method generally produces a network with superfluous elements.

A simpler structure which possesses the same terminal properties as the two port of Fig. 2 is shown in Fig. 3. This was proposed by P. Penfield of MIT. A one port which is defined by $V = I = 0$ is easily deduced from Fig. 3(a) and 3(b) and is shown in Fig. 3(c) and 3(d). It is easy to verify that these networks are as prescribed. Thus, as shown in Fig. 3(c), the $-1\,\Omega$ load imposes the condition $v_{r2} = 0$. Similarly, at the $+1\,\Omega$ load, $v_{i3} = 0$. Since the effect of the circulator is to make $v_{i1} = v_{r2}$, $v_{r1} = v_{i3}$, we have $v = v_{i1} + v_{r1} = 0$, $i = v_{i1} - v_{r1} = 0$. v_i and v_r are incident and reflected voltages.

The networks of Figs. 2 and 4 are certainly peculiar, but an additional anomaly for the one port defined by

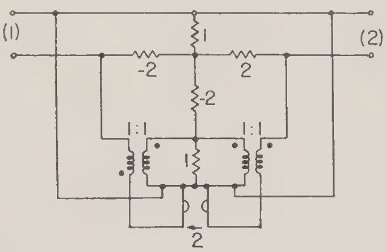

Fig. 2—Realization of

$$[A, B] = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{bmatrix}.$$

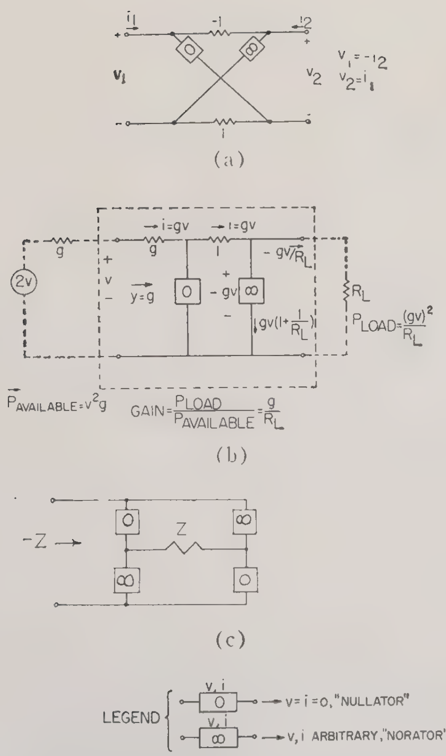

Fig. 3—Equivalent circuits using "nullator" ($v = i = 0$), and "norator" (v, i, arbitrary). (a) Gyrator, (b) unilateral voltage amplifier, (c) negative impedance converter.

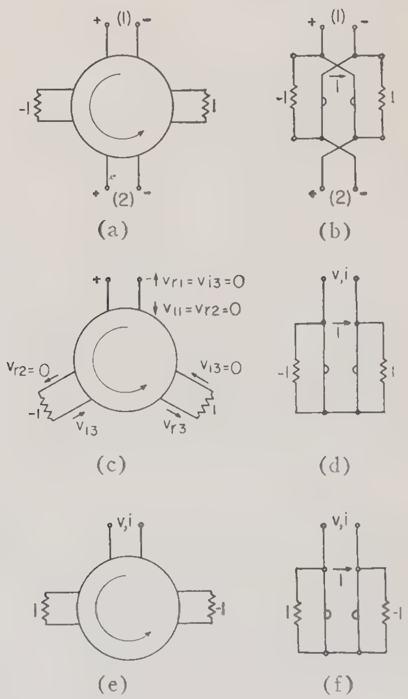

Fig. 4—Degenerate one-port and two-port networks.

(a), (b): $\begin{bmatrix} 1 & 0 \\ 0 & 0 \end{bmatrix}\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} 0 & 0 \\ 1 & 0 \end{bmatrix}\begin{bmatrix} i_1 \\ i_2 \end{bmatrix}$.

(c), (d) $v = i = 0$ ("nullator"), (e), (f) v and i arbitrary ("norator").

$v=i=0$ is that all the structures shown use nonreciprocal elements, lossy elements, and active elements, yet the defining equations imply a reciprocal lossless network (since input power $=vi=0$). It has, in fact, been shown[3] that despite its lossless, reciprocal terminal behavior, the one port defined by $v=i=0$ requires nonpassive, nonreciprocal elements for its realization.

A further property of the networks of Fig. 2 and Fig. 4(a) and (b) is that when port 1 (the port at which $v=i=0$) is terminated, say in an open or short circuit, the voltage and current at port 2 are completely arbitrary. A simple equivalent circuit for this one port is shown in Fig. 4(e) and (f). This kind of anomalous structure is the basis for the synthesis of an $[A, B]$ matrix whose normal rank is $r \leq 2n$. A further, most important property of the v, i arbitrary element is that it defines the only possible linear nonreciprocal one port.[4] This element, together with the $v=i=0$ element, consititute two basic circuit building blocks for which the following names are here proposed: the null element, ($v=i=0$) to be called the "nullator," the nonreciprocal one port (v, i arbitrary) to be called the "norator." Fig. 3 demonstrates interconnections of the nullator and norator to form a gyrator, a unilateral voltage amplifier, and a negative-impedance converter.

At this point let us consider the synthesis problem for a general n port when an $[A, B]$ representation exists but no special restrictions are placed on the normal rank of this matrix. A and B are each rectangular, real, rational $m \times n$ matrices.[5] To handle this case $[A, B]$ will be transformed into a simple canonic form. Thus, consider:

$$[A, B]\begin{bmatrix} V \\ I \end{bmatrix} = Q[A, B]PP^{-1}\begin{bmatrix} V \\ I \end{bmatrix}$$

$$= [A', B']\begin{bmatrix} V' \\ I' \end{bmatrix} = 0 \qquad (11)$$

with

$$[A', B'] = Q[A, B]P \quad \begin{bmatrix} V' \\ I' \end{bmatrix} = P^{-1}\begin{bmatrix} V \\ I \end{bmatrix}, \qquad (12a)$$

where Q is $m \times m$, P is $2n \times 2n$ and these are determined[6] to give the canonic form

$$[A', B'] = Q[A, B]P, = \begin{bmatrix} E_r & 0_{r,n-r} \\ 0_{m-r,r} & 0_{m-r,n-r} \end{bmatrix}, \qquad (12b)$$

[3] D. Youla, "Internal Structure of One Ports Simultaneously Open and Short Circuited," Polytechnic Institute of Brooklyn, Brooklyn, N. Y., MRI Memorandum No. 47; February, 1961.

[4] There are only three possible linear one-ports corresponding to $r=0, 1, 2$. The nullator has $r=2$, the norator $r=0$, and the $r=1$ network has an admittance or impedance. Using the Lorenz reciprocity relations, only the norator is nonreciprocal.

[5] Using non-square matrices the $v=i=0$ one-port may be written in $[A, B]$ form as

$$\begin{bmatrix} 1 \\ 0 \end{bmatrix} v_1 = \begin{bmatrix} 0 \\ 1 \end{bmatrix} i_1.$$

[6] Aitken, *op. cit.*, pp. 66–67. This reference applies to matrices of constants, but the procedure readily extends to the rational case.

Q and P represent matrices of real, rational functions of p which perform only elementary row and column operations on $[A, B]$; hence Q and P are nonsingular qn of normal ranks m and $2n$ respectively. The resultant current and voltage vectors V', I' satisfy the last equality of (11). Hence referring to (12b), we may define the individual voltages and currents of the transformed network and distinguish two cases:

Case a:

$$0 \leq r < n$$
$$v_1' = v_2' = \cdots = v_r' = 0,$$

all other voltages and currents are arbitrary.

Case b:

$$n \leq r < 2n$$
$$v_1' = v_2' = \cdots = v_n' = 0$$
$$i_1' = i_2' = \cdots = i_{r-n} = 0,$$

the remaining $2n-r$ currents are arbitrary.

In Case a, the network represented by $[A', B']$ consists of n decoupled one ports. Of these, r are short circuits and $n-r$ have arbitrary voltages and currents at their ports so that each of the latter can be represented by the norator of Fig. 4(f).

In Case b, $[A', B']$ is again represented by n decoupled one ports, $r-n$ of which are nullators as in Fig. 4(e) ($v_k'=i_k'=0$), while the remaining $2n-r$ are short circuits.

In either case, the n-decoupled one ports are connected across n of the ports of the $2n$ port representing P^{-1}, and the remaining n ports of P^{-1} constitute the input terminal pairs for the prescribed V, I variables. It remains to be shown that the P^{-1} transformation can always be realized as a $2n$-port network.

Denote the matrix

$$P^{-1} = \begin{bmatrix} A_1 & A_2 \\ A_3 & A_4 \end{bmatrix}.$$

Then if the second equation of (12a) is written in $[A, B]$ form,

$$\begin{bmatrix} A_1 & -E_n & A_2 & 0 \\ A_3 & 0 & A_4 & -E_n \end{bmatrix}\begin{bmatrix} V \\ V' \\ I \\ I' \end{bmatrix} = 0. \qquad (13)$$

The minor

$$\begin{vmatrix} -E_n & 0 \\ 0 & -E_n \end{vmatrix}$$

in (13) forms a nonvanishing $2n$th order determinant, and hence, by Theorem 2, the network corresponding to P^{-1} can be realized from its $[A, B]$ representation.

We may summarize the results of this section by the following general theorem:

Theorem 3:

If an $m \times 2n$ $[A, B]$ matrix ((*i.e.*, A and B rectangular and $m \times n$) of normal rank $r \leq 2n$ containing real, rational functions is prescribed, the description $AV = BI$ may always be realized as a lumped n-port network containing passive elements and negative resistors.[7]

It might be reiterated at this point that a necessary condition on an $[A, B]$ matrix for realization as an n port with passive elements is that $A+B$ be of normal rank n. If this is not satisfied negative resistors are required and one of the methods described above may be used.

Synthesis with a Restricted Number of Resistors

A. Networks with Only Lossless Elements and Negative Resistors

In the preceding section, it was shown that even in the absence of any usual formalism such as immittance or scattering, the synthesis of any linear, lumped system can be obtained by recourse to the $[A, B]$ formalism which always exists when a network has a graph. In this section, special attention is devoted to the class of problems in which an immittance representation *does* exist and a minimal resistor element synthesis is presented for this case. A final result is that any rational $n \times n$ immittance matrix whatever, with real coefficients, can always be constructed as an n port in which passive lossless elements are used and which in addition contains at most n positive and n negative resistors. Again, the presentation proceeds from restrictive cases to the most general result. Theorem 4 considers n-port networks which have only lossless lumped elements and negative resistors. Theorem 5 then extends the synthesis to rational matrices with real coefficients whose left-half plane and right-half plane poles are arbitrary in location and order, but whose boundary poles ($p = j\omega$) are simple. The problem of competely unrestricted poles is then treated, first for a one port and finally for an n port. Theorem 6 considers the special one port which has an immittance function odd in the complex-frequency variable p; and Theorem 7 generalizes this result to the one port with a completely arbitrary rational immittance function with real coefficients, and it is shown that 1 positive and 1 negative resistor suffice for the general one port. The final generalization to $n \times n$ immittance matrices containing completely arbitrary rational functions (*i.e.*, no restrictions on pole locations or multiplicity) with real coefficients is given in Theorem 8, which states that, at most, n positive and n negative resistors are required for an n port. Theorem 9 states that if the rational $n \times n$ immittance matrix is symmetric, then the realization only requires reciprocal elements, but again n positive and n negative resistors are sufficient.

[7] Referring to Fig. 3, the nullator and norator may be used as building blocks instead of the gyrator and negative resistor.

Consider first a network, all of whose resistors are negative. It is clear that the power absorbed in such a structure must always be negative so that at real frequencies, since

$$2P(\omega) = I^*(Z(p) + Z'(-p))I\big|_{p=j\omega} \leq 0$$

then

$$Z_H = Z(p) + Z'(-p),$$

must be the matrix of a negative definite (or semidefinite) form on $j\omega$. (In the above equations prime ($'$) indicates matrix transpose, asterisk (*) indicates complex conjugate transpose, and complex frequency is $p = \sigma + j\omega$.) Thus $-Z(p)$ for such a network must have a positive power form on $j\omega$. Even more strongly, it can be shown that a necessary condition for a network, $Z(p)$, to contain only negative resistors is that $-Z(-p)$ be a positive real matrix. To demonstrate this, suppose that we visualize the n port $Z(p)$ as a reactive $n+m$ port $Z_0(p)$ network with all m resistors accounted for by being connected to m ports of $Z_0(p)$, and with the remaining n ports considered as the input terminal pairs of $Z(p)$. This is shown in Fig. 5. If $Z_0(p)$ is partitioned to make evident the m and n ports, we have

$$Z_0(p) = \begin{bmatrix} Z_{nn}(p) & Z_{nm}(p) \\ Z_{mn}(p) & Z_{mm}(p) \end{bmatrix}.$$

The elements here are submatrices and represent the self- and mutual impedances between the n and m ports.

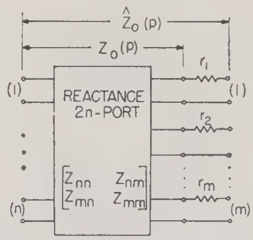

Fig. 5—Reactive $n+m$ port with resistive terminations.

If the resistors are connected at the m ports, but for the moment these ports are not closed, then the matrix $Z_0(p)$ becomes

$$\hat{Z}_0(p) = \begin{bmatrix} Z_{nn}(p) & Z_{nm}(p) \\ Z_{mn}(p) & Z_{mm}(p) - R \end{bmatrix}, \tag{14}$$

where $-R = -\text{diag}\ (r_1, \cdots, r_m)$, $r_k > 0$, and lists all m negative resistors of the network. If, now, ports m are closed around the resistors $-r_k$, the impedance matrix measured at the n accessible ports is $Z(p)$. The reduction formula [8] may be used to get an expression for $Z(p)$ in terms of the elements of (14):

$$Z(p) = Z_{nn}(p) - Z_{nm}(p)[Z_{mm}(p) - R]^{-1}Z_{mn}(p). \tag{15}$$

Observe that since $Z_0(p)$ is reactive, it must be skew-hermitian[8] on $p=j\omega$, or in terms of the complex-frequency variable p,

$$Z_{nn}(p) = -Z_{nn}'(-p), \quad Z_{mm}(p) = -Z_{mm}'(-p), \quad Z_{nm}(p) = -Z_{mn}'(-p), \tag{16}$$

everywhere in the complex p plane.

Substituting (16) into (15) gives

$$Z(p) = -Z_{nn}'(-p) + Z_{mn}'(-p)[Z_{mm}'(-p)+R]^{-1}Z_{nm}'(-p) = -[Z_{nn}(-p) - Z_{nm}(-p)[Z_{mm}(-p)+R]^{-1}Z_{mn}(-p)]', \tag{17}$$

and

$$-Z'(-p) = Z_{nn}(p) - Z_{nm}(p)[Z_{mm}(p) + R]^{-1}Z_{mn}(p). \tag{18}$$

Comparing (18) with (15), it is clear that $-Z'(-p)$ is the impedance matrix of a reactive $n+m$ port network, m ports of which are terminated in *positive* resistors r_k. Thus $-Z'(-p)$ is a PR (positive real) matrix and its transpose $-Z(-p)$ is also PR. This establishes necessary conditions for a negative-resistance network.

A sufficiency proof is demonstrated if, given an $n\times n$ $Z(p)$, such that $-Z(-p)$ is a PR matrix, it can be shown that a network corresponding to $Z(p)$ exists containing only pure reactive elements and negative resistors. It can be shown that such a network is easy to synthesize. Since $-Z(-p)$ is PR, so too is $-Z'(-p)$, and this may be constructed as a reactive $2n$ port, n of whose ports are terminated in positive resistors [9]. If the reactive $2n$ port is $Z_0(p)$, then in terms of the self- and mutual impedances at the two sets of network ports (n accessible, n terminated in positive resistors), the form of $-Z'(-p)$ is given by (18). Now if the positive resistors r_k are each changed to $-r_k$, it is clear that (18) changes into (15), and we have exhibited $Z(p)$ by merely synthesizing $-Z'(-p)$ and reversing the signs of all the n positive resistors. We therefore have the following theorem:

Theorem 4:

> The necessary and sufficient condition that an $n\times n$ $Z(p)$ of real, rational functions corresponds to a network containing only lossless elements and negative resistors is that $-Z(-p)$ be a PR matrix. Alternately the scattering matrix $S(p)$ corresponding to $Z(p)$ satisfies the requirement that $S^{-1}(-p)$ is a bounded real scattering matrix. Further, the network representing $Z(p)$ may always be synthesized as a structure containing at most n[9] negative resistors.

As a consequence of Theorem 4, it is simple to show that any rational, real $Z(p)$ of dimensions $n\times n$ may be realized with at most n positive and n negative resistors provided that $Z(p)$ has no boundary poles other than simple ones with nonnegative residue matrices. To see this, represent $Z(p)$ as

$$Z(p) = Z_R(p) + Z_L(p) + Z_B(p),$$

where the three impedance terms respectively contain the right-half plane, left-half plane and boundary poles of $Z(p)$. The term $Z_B(p)$ is easily disposed of since this contains only simple boundary poles with nonnegative residue matrices and hence merely represents a purely reactive network in series with the other two. (Higher order poles on the boundary including the point at infinity are excluded by hypothesis.)

The terms $Z_R(p)$ and $Z_L(p)$ are both finite everywhere on $j\omega$ since they contain no boundary poles, hence $R=\mathrm{diag}(r_1, r_2, \cdots, r_n)$, $r_k\geq 0$ can always be chosen so that the hermitian part of $Z_1=-R+Z_R(p)$ is the matrix of a negative definite form and the hermitian part of $Z_2=R+Z_L(p)$ is the matrix of a positive definite form. In addition, $Z_2(p)$ is analytic in the right-half plane, hence it is a PR matrix and is realizable with reactive elements and n positive resistors. Further, since $Z_R(p)$ has only right-half plane poles, $Z_R(-p)$ is analytic in the right-half plane and with

$$-Z_1(-p) = R - Z_R(-p),$$

it is clear that $-Z_1(-p)$ is PR since it has a positive hermitian part, no poles on $j\omega$, and is analytic in the right-half plane. Thus by Theorem 4, $Z_1(p)$ is realizable with lossless elements and n negative resistors. Since $Z(p)=Z_1(p)+Z_2(p)+Z_B(p)$, we have synthesized $Z(p)$ with n positive and n negative resistors and thus there follows Theorem 5.

Theorem 5:

> Any $n\times n$ rational, real immittance matrix with unrestricted poles, except that those on the boundary are simple with nonnegative residue matrices, can be realized as a network containing lossless elements and at most n positive and n negative resistors.

This theorem indicates that under special restrictions (simple boundary poles), synthesis with a minimum number of resistors is possible. It is clear that in the special case that $Z(p)$ is symmetric, so too will be $Z_1(p)$, $Z_2(p)$, $Z_B(p)$, and thus, these components are realized with reciprocal networks. Hence a corollary of Theorem 5 is that under the constraint of simple boundary poles with nonnegative residue matrices at these poles, an otherwise arbitrary $n\times n$ *symmetric* rational immittance matrix with real coefficients can be synthesized as an n port with *reciprocal* lossless elements and at most n-positive and n-negative resistors.

We now proceed to the general case of an immittance matrix in which arbitrary-order boundary poles (including the point at infinity) are permitted. First the one-port or driving-point impedance is considered, and this case is treated in the following section.

[8] All discussions pertain to general nonreciprocal networks.

[9] It is readily shown that if the rank of $Z_H(p)=Z(p)+Z(-p)$ is $r\leq n$, then at most r resistors are required.

B. General Driving-Point Immittance Functions

In this section the synthesis of an arbitrary, rational, real, driving-point immittance function will be discussed and it will be shown that reciprocal elements and at most one negative resistor are needed. Consider first a passive reciprocal two-port network terminated in a -1-ohm resistor. The input reflection factor to such a network is

$$s_0 = s_{11} + \frac{s_{12}^2 s}{s_{11} - s s_{22}}, \tag{19}$$

where s is the reflection factor of the terminating impedance, and S_{jk} are two-port scattering functions. When the termination approaches -1, $s \to \infty$, so that (19) becomes

$$s_0 = s_{11} - \frac{s_{12}^2}{s_{22}} \cdot \tag{20}$$

If we set $s_{11} = 0$

$$s_0 = - \frac{s_{12}^2}{s_{22}} \bigg|_{s_{11}=0} \cdot \tag{21}$$

We shall now show that a symmetric bounded real scattering matrix (hence reciprocal passive network) defined by appropriate functions $s_{11}=0$, s_{12}, s_{22} corresponds to any prescribed, arbitrary, real, rational function s_0. We obtain s_0 by terminating the passive two port in -1 ohm.

Write the prescribed reflection factor $s_0(p)$ in the form of numerator and denominator polynomials

$$s_0(p) = \frac{N(p)}{D(p)} = \frac{N_R(p) N_L(p)}{D_R(p) D_L(p)}, \tag{22}$$

where the subscripts R and L designate right- and left-half plane root factors respectively, and any roots on the finite $p = j\omega$ axis are included in the "R" factors. Thus

$$s_0 = \frac{N_R(p)}{D_L(p)} \cdot \frac{1}{\dfrac{D_R(p)}{N_L(p)}}$$

$$= K^2 \frac{N_R(p)\overline{N}_R(p)}{D_L(p)\overline{D}_L(p)} \cdot \frac{1}{K^2 \dfrac{D_R(p)\overline{N}_R(p)}{N_L(p)\overline{D}_L(p)}} \cdot \tag{23}$$

K^2 is a real, positive constant and the polynomials $\overline{N}_R(p)$ and $\overline{D}_L(p)$ are chosen to make $N_R\overline{N}_R$, and $D_L\overline{D}_L$ perfect squares. Furthermore, the surplus factors of $\overline{N}_R$ and $\overline{D}_L$ are chosen of sufficiently high order so that

$$\text{Degree } (D_L\overline{D}_L) \geq \text{Degree } (N_R\overline{N}_R), \tag{24a}$$

$$\text{Degree } (N_L\overline{D}_L) \geq \text{Degree } (D_R\overline{N}_R). \tag{25a}$$

Note that if D_L and/or N_L is a constant, we arbitrarily insert a left-half plane root factor located on the σ axis and place this within $\overline{D}_L(p)$.

Using (21) as a guide, we now identify

$$s_{12}^2 = K^2 \frac{N_R(p)\overline{N}_R(p)}{D_L(p)\overline{D}_L(p)} = K^2 \frac{N_{12}^2(p)}{D_{12}^2(p)}, \tag{26}$$

$$s_{12} = K \frac{N_{12}(p)}{D_{12}(p)}, \tag{27}$$

$$s_{22} = -K^2 \frac{D_R(p)\overline{N}_R(p)}{N_L(p)\overline{D}_L(p)} \cdot \tag{28}$$

Observe that s_{12} and s_{22} are analytic in the right-half plane and everywhere on the boundary, including the point at infinity by virtue of the method outlined for choosing the surplus factors. Also, (21) is satisfied. We now choose K in a manner which guarantees the passivity of the 2 port whose scattering matrix is

$$S = \begin{bmatrix} 0 & s_{12} \\ s_{12} & s_{22} \end{bmatrix}.$$

This is done in the following manner. If S is bounded real, its elements must be analytic in the right-half plane (already satisfied), and the hermitian matrix

$$E - S(-p)S(p) \big|_{p=j\omega} = Q(\omega^2) \tag{29}$$

must be positive; that is

$$|s_{12}|^2 = s_{12}(p)s_{12}(-p) \big|_{p=j\omega} \leq 1, \tag{30}$$

$$|s_{22}|^2 = s_{22}(p)s_{22}(-p) \big|_{p=j\omega} \leq 1, \tag{31}$$

and

$$\text{Det } Q(\omega^2) \geq 0. \tag{32a}$$

But

$$\text{Det } Q(\omega^2) = [1 - |s_{12}(j\omega)^2|]^2 - |s_{22}(j\omega)|^2 \geq 0, \tag{32b}$$

or

$$|s_{12}(j\omega)|^2 + |s_{22}(j\omega)| \leq 1. \tag{33}$$

If this is true, (30) and (31) are automatically satisfied, so that only (33) is needed to assure the positive character of Q.

Substituting (27) and (28) into (33), we now have an equation which defines K^2:

$$K^2 \left[\left| \frac{N_{12}}{D_{12}} \right|^2 + \left| \frac{D_R\overline{N}_R}{N_L\overline{D}_L} \right| \right]_{p=j\omega} \leq 1. \tag{34}$$

Since $D_{12}(p)$, $N_L(p)$, $\overline{D}_L(p)$ have no poles on $j\omega$, the quantities inside the bracket have a finite maximum M, and K^2 is chosen as $1/M$, so that inequality (34) can always be satisfied. This completes the determination of S, and the prescribed input immittance is obtained by synthesizing the two-port corresponding to S and then terminating the output of this two-port in a -1 ohm resistor.

A special case is of particular importance. Suppose the prescribed driving-point impedance is an odd function. Then

$$s_0(p)s_0(-p) = \frac{Z(p)-1}{Z(p)+1}\;\frac{-Z(p)-1}{-Z(p)+1} = 1,$$

or $s_0(p)$ is an all-pass function with no boundary poles and may be written as a finite Blaschke product.

$$S_0(p) = \Pi\,\frac{(p-\alpha_k)(p-\alpha_k^*)}{(p+\alpha_k)(p+\alpha_k^*)}$$

where the α_k may be either in the right- or left-half plane.

Since right- and left-half plane roots are always paired, $s_{12}(p)$ and $s_{22}(p)$ [given in (27) and (28)] similarly contain matched root and pole pairs, and the amplitudes of these functions are respectively K and K^2 on $p=j\omega$. Thus (34) becomes

$$2K^2 \leq 1,$$

and the maximum permissible K^2 is given by

$$K^2 = 1/2.$$

Referring to (32), this means

$$\operatorname{Det} Q(\omega^2) = 0,$$

so that the rank of Q is 1. $S(p)$ therefore corresponds to a *reduced*[10] network in the sense of Gewertz [10] and hence only requires one positive resistor for its synthesis [9]. We therefore have the following theorem.

Theorem 6:

> A real, odd, rational-driving point immittance function $[z(p)=-z(-p)]$ may be synthesized with lossless reciprocal lumped elements and at most one positive and one negative resistor. Alternately this is true when the input reflection function is real and rational and satisfies $s_0(p)s_0(-p)=1$.

As an example of the application of Theorem 6, let us indicate the synthesis of $Z(p)=p^3$. The reflection factor is:

$$s_0 = \frac{p^3-1}{p^3+1} = \frac{(p-1)(p^2+p+1)}{(p+1)(p^2-p+1)},$$

which, according to (23), we write in the form

$$s_0 = K^2\frac{(p-1)(p-1)}{(p+1)(p+1)}\;\frac{1}{K^2\dfrac{p^2-p+1}{p^2+p+1}\cdot\dfrac{p-1}{p+1}}\,.$$

[10] A "reduced" network in the sense of Gewertz for a reciprocal 2 port is one in which rank [Re $Z(j\omega)$] = 1. This condition is satisfied if $Q(\omega^2)$ is of rank 1.

Then

$$s_{11} = 0;\qquad s_{12} = s_{21} = K\,\frac{p-1}{p+1};$$

$$s_{22} = -K^2\frac{p^2-p+1}{p^2+p+1}\;\frac{p-1}{p+1},$$

and with

$$K = 1/\sqrt{2},$$

these functions define a physically realizable two port containing one positive resistor which is realizable by the methods of Oono [9] or Gewertz [10]. $Z(p)$ is obtained by terminating the output of this two port with a $-1\ \Omega$ resistor.

We will now generalize Theorem 6 and show that *any* real rational driving-point immittance function requires at most one positive and one negative resistor for its synthesis as a reciprocal network. The technique to be used requires the removal of boundary poles and then a synthesis according to Theorem 5.

Consider first a prescribed immittance function $Z(p)=N(p)/D(p)$ which is not purely odd. Now augment $Z(p)$ to get

$$\overline{Z}(p) = \frac{N(p)}{D(p)} - Ap - \frac{B}{p}\,\cdot \tag{35}$$

We will choose A and B (real, positive constants) so that $Z(p)$ has no zeros anywhere on $p=j\omega$, including the point at infinity. To see how this is done, rewrite (35) as

$$\overline{Z}(p) = \frac{pN - Ap^2D - BD}{pD}, \tag{36}$$

Now examining (36),

1) choose A so that if p^2D is of the same degree as pN, the highest powers of pN and Ap^2D do not cancel. This insures that the denominator degree of $\overline{Z}(p)$ does not exceed the numerator degree. Eq. (35) shows that whenever N/D has a zero at a finite $p_k=j\omega_k$, this zero does not appear in $\overline{Z}(p)$ provided that

2) we choose A and B so that $Ap+B/p$ does not vanish at any of the ω_k. Finally, in order that no new boundary zeros be created in $\overline{Z}(p)$, we must consider the points at which $Z(p)$ has a zero real part. These will be finite in number since $Z(p)$ is initially prescribed as not purely odd, *i.e.*, its real part is not always zero on $j\omega$.

3) We then choose A and B so that at these pure reactance points of $Z(j\omega)$, the reactance of $Ap+B/p$ does not cancel the reactance of $Z(j\omega)$.

There are then a finite number of forbidden values of A, B given by 1)–3) that must be avoided. If the constants A and B are chosen to avoid these forbidden values, $\overline{Z}(p)$ has no zeros on $j\omega$ and $1/\overline{Z}(p)$ has no boundary poles anywhere (including the point at infinity). We may then synthesize $1/\overline{Z}(p)=\overline{Y}(p)$ by separating it into the sum of two parts, one of which contains the left-half plane poles and the other of which contains the right-half plane poles. As indicated in Theorem 5, this synthesis only requires one positive and one negative resistor. Finally, $Z(p)$ is constructed by adding in series to the network corresponding to $\overline{Z}(p)$ the pure reactance $Ap+B/p$. The only case excluded by this process is that in which $Z(p)$ is purely odd but this is taken care of by Theorem 6. We have therefore arrived at the following result.

Theorem 7:

> Any arbitrary real, rational driving point immittance function whose zeros and poles are completely unrestricted as to multiplicity and location in the complex p plane may be realized as a lumped network consisting of reciprocal lossless elements and at most one positive and one negative resistor.[11]

C. Synthesis of n Ports With n-Positive and n-Negative Resistors

In this section we extend Theorem 7 to the most general n port and establish a synthesis with n-positive and n-negative resistors, using a method of proof which parallels that already given for a one port.

Consider then an arbitrary $n\times n$ $Z(p)$, which is generally non-reciprocal and contains real, rational elements but which is not skew-symmetric, *i.e.*, $Z(p) \neq -Z'(-p)$. Let us parallel the technique of the previous section and first augment $Z(p)$ to remove boundary zeros.

$$\overline{Z}(p) = Z(p) - \overline{X}(p), \tag{37}$$

where the elements of $\overline{X}(p)$ are given by

$$\overline{X}_{ii} = A_i p + B_i/p; \qquad \overline{X}_{ij} = Ap + B/p + \alpha;$$
$$\overline{X}_{ji} = Ap + B/p - \alpha. \tag{38}$$

$$A_i,\ B_i,\ A,\ B,\ \alpha \text{ real, positive;} \qquad A_i \geq A,\ B_i \geq B.$$

The matrix $\overline{X}(p)$ is physically realizable as a pure reactance n port since it is skew-symmetric on $p=j\omega$, its elements are analytic in Re $p>0$, and all boundary poles are simple with positive residue matrices [5].

The inverse of $\overline{Z}(p)$ has elements

$$\bar{y}_{ij}(p) = \frac{A_{ji}(p)}{\bar{\Delta}(p)}, \tag{39}$$

where A_{ji} is the cofactor of $\bar{z}_{ij}$, an element of $\overline{Z}(p)$, and $\bar{\Delta}(p)$ is the determinant of $\overline{Z}(p)$. The numbers $q_i=(A_i, B_i, A, B, \alpha)$ are to be chosen according to (38) and must avoid a finite set of forbidden values. The choice is to be made so that $\bar{y}_{ij}(p)$ has no boundary poles. Appendix II shows that this may always be done provided $Z(p)\neq -Z'(-p)$. $\overline{Z}(p)$ is then synthesized according to Theorem 5, and $Z(p)$ finally constructed by adding in series with $\overline{Z}(p)$ the reactance n port for $\overline{X}(p)$.

To obtain the final general result we must treat the remaining case in which an immittance matrix which is skew-symmetric is prescribed, $Z(p)=-Z'(-p)$ in this case and

$$S(p)S'(p) = (Z(-p) - 1)(Z(p) + 1)^{-1}(-Z(p) - 1) \cdot(-Z(p) + 1)^{-1} = E,$$

and $S(p)$ is para-unitary.[12] Since this is true, $S(p)$ contains no boundary poles.[13] We may therefore write

$$S(p) = g_1(p)S(p)g^{-1}(p)EE, \tag{40}$$

where $g_1(p)$ is a scalar Blaschke product, analytic in the right-half plane, whose numerator contains all the right-half plane pole factors of $S(p)$, so that g_1S has all its right-half p-plane poles cancelled. Thus with

$$U_1 = g_1S, \qquad U_2 = E; \qquad \Lambda = g^{-1}E, \tag{41a}$$

we have

$$S(p) = U_1\Lambda U_2. \tag{41b}$$

This is a para-unitary transformation of the diagonal para-unitary matrix Λ. Since U_1 and U_2 are para-unitary and analytic in the right-half p plane, (41b) corresponds to a $2n$-port lossless network whose scattering matrix is

$$T = \begin{bmatrix} 0 & g_1S \\ E & 0 \end{bmatrix}, \tag{42}$$

and which has connected to its n-output ports the n-port Λ [9]. This latter structure, in turn, consists of n-decoupled one ports, each of whose reflection factors is $g^{-1}(p)$; and since $g^{-1}(p)g^{-1}(-p)=1$, this one port is synthesized according to Theorem 6 with lossless reciprocal elements and a single positive and negative resistor. Thus the network for S is exhibited at the remaining n ports of T and contains lossless elements and at most

[11] The authors gratefully acknowledge the key suggestion of Mr. Jack Sipress of Bell Telephone Labs. to the effect that the boundary zeros of $Z(p)$ may be removed by adding $-(Ap+B/p)$.

[12] A para-unitary matrix is unitary for $p=j\omega$. That is,

$$U_1(p)U_2'(-p) = E.$$

[13] The para-unitary requirement gives $\sum |s_{ij}(j\omega)|^2=1$ and thus if any s_{ij} contained a $p=j\omega$ pole it could not be cancelled by any other term since each element of the sum is positive, hence such a pole is excluded.

n-positive and n-negative resistors. A schematic diagram of the network synthesis is shown in Fig. 6. The theorem which summarizes this most general result is therefore as follows.

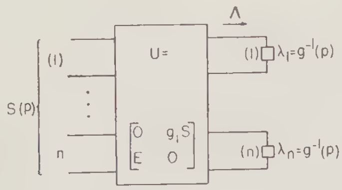

Fig. 6—Synthesis of a para-unitary scattering matrix, $S(p)S'(-p) = E$.

Theorem 8:

To any $n \times n$ immittance or scattering matrix[14] (not necessarily symmetric) of real rational functions with arbitrary order and location of zeros and poles, there corresponds an n-port network composed of lossless elements and at most n-positive and n-negative resistors.

It is interesting to point out that when the synthesis technique just described is applied to a symmetric immittance matrix, the structure that results uses only reciprocal elements. This follows, since in the symmetric case, it is only necessary to augment $Z(p)$ with a symmetric $\overline{X}(p)$ rather than a skew-symmetric reactance matrix in order to obtain a $\overline{Y}(p)$ without boundary poles. As a matter of fact, there is never any need for skew-symmetric $\overline{X}_{ij}$ terms as far as the poles of A_{ji} are concerned. The necessity for skew-symmetry of $\overline{X}$ is to take care of the case where all the z_{ij} are odd, but $Z(p)$ is neither symmetric nor skew-symmetric. If this occurs, $\Delta(p)$ is either purely even or odd; but by augmenting with a skew-symmetric $\overline{X}(p)$, we obtain a $\overline{\Delta}(p)$ which is neither even nor odd, and the boundary zeros of this can be removed. If $Z(p)$ is symmetric, however, it will then always be possible to eliminate the boundary zeros of $\Delta(p)$ by using a symmetric $\overline{X}(p)$ (hence a series connected reciprocal lossless network), unless $Z(p)$ is both purely odd and symmetric, in which case Δ and $\overline{\Delta}$ are both either even or odd. Such a situation again requires separate treatment, and we now wish to show that if $Z(p) = -Z(-p)$ with $Z(p) = Z'(p)$ then a suitable *reciprocal* transformation leads to an appropriate synthesis.

It has been shown[15] that if $S(p)$ is symmetric and para-unitary, it can be reduced to para-unitary diagonal form by a reciprocal para-unitary transformation given by

$$\Lambda = USU', \tag{43}$$

with U para-unitary.

[14] The presumption is that both these representations simultaneously exist.

[15] Oono and Yasuura, *op. cit.*, see Theorem 17, p. 151.

We can consider that S in (43) corresponds to the prescribed odd symmetric $Z(p)$ and hence

$$S = W\Lambda W', \tag{44}$$

with $W = U^{-1}$.

As before, we remove the right-half plane poles of W with a scalar Blaschke product which leaves the para-unitary property unchanged:

$$S = (gW)\left(\frac{1}{g^2}\Lambda\right)(W'g) = \overline{W}\overline{\Lambda}\overline{W}', \tag{45}$$

where $\overline{W} = gW$ is para-unitary and analytic in the right-half plane. The transformation of (45) therefore corresponds to a reciprocal lossless $2n$ port [9] with scattering matrix

$$T = \begin{bmatrix} 0 & \overline{W} \\ \overline{W}' & 0 \end{bmatrix}.$$

At each of the n ports of T, a reciprocal one port corresponding to the diagonal elements of $\overline{\Lambda}$ is connected and synthesized with one positive and one negative resistor according to Theorem 6. That is, since $\overline{\Lambda}$ is diagonal and para-unitary, each of its elements is a Blaschke product representing the input reflection factor of a one port. The remaining n ports of T exhibit the required network for S. The following theorem may therefore be stated.

Theorem 9:

Any $n \times n$ *symmetric* immittance or scattering matrix containing real, rational functions, but with poles and zeros of arbitrary order and location, may be synthesized by a reciprocal network containing lossless elements and at most n-positive and n-negative resistors.

Appendix I

Transformation of an [A, B] Matrix

We show here that if

$$[A, B]\begin{bmatrix} V \\ I \end{bmatrix} = [A, B]PP^{-1}\begin{bmatrix} V \\ I \end{bmatrix} = [A', B']\begin{bmatrix} V' \\ I' \end{bmatrix},$$

with P a $2n \times 2n$ permuting matrix of constants

$$\begin{bmatrix} V \\ I \end{bmatrix} = P\begin{bmatrix} V' \\ I' \end{bmatrix},$$

then P has a $2n$-port representation using ideal transformers, gyrators, and $\pm R$. To do this, the interchange of only 2 columns need be considered, since P may be taken as the product of a multiplicity of permutation matrices corresponding to the cascade of a set of $2n$-port networks, each of which is associated with one column interchange. The permutation by P of the ith column of A with the ith column of B is represented by a gyrator at port i which merely interchanges the voltage and current at that port. Now consider the

interchange of *any* two columns in A, B as represented by (9a). Partition the $2n\times 2n$ matrix P into four $n\times n$ submatrices,

$$P = \begin{bmatrix} P_1 & P_2 \\ P_3 & P_4 \end{bmatrix}. \tag{46}$$

The matrix P differs from the identity matrix E_{2n} only in that if the ith and jth columns of $[A, B]$ are to be permuted, then the ith and jth columns of E_{2n} are exchanged to form P. From (9a) and (46),

$$V = P_1V' + P_2I',$$
$$I = P_3V' + P_4I',$$

and putting this in $[A, B]$ form

$$\begin{bmatrix} E_1 & -P_1 \\ 0_n & P_3 \end{bmatrix}\begin{bmatrix} V \\ V' \end{bmatrix} = \begin{bmatrix} 0_n & P_2 \\ E_2 & -P_4 \end{bmatrix}\begin{bmatrix} I \\ I' \end{bmatrix}. \tag{47a}$$

E_1 and E_2 are n-rowed identity matrices with subscripts used merely to identify their location in (47a). It must now be shown that

$$[A_P, B_P] = \begin{bmatrix} E_1 & -P_1 & 0_n & P_2 \\ 0_n & P_3 & E_2 & -P_4 \end{bmatrix}, \tag{47b}$$

which is clearly of rank $2n$, has a non-vanishing complementary $2n$-columned minor. Suppose that the interchanged two columns of P occur in

$$\begin{bmatrix} -P_1 \\ P_3 \end{bmatrix};$$

then E_1 and $-P_4$ form a complementary nonvanishing minor, since $-P_4 = -E_n$. A similar situation occurs when the interchanged columns of P occur in

$$\begin{bmatrix} P_2 \\ -P_4 \end{bmatrix},$$

for then E_1 and P_3 form the non-zero minor.

The final case to consider is when the ith column of

$$\begin{bmatrix} -P_1 \\ P_3 \end{bmatrix}$$

has been interchanged with the jth column of

$$\begin{bmatrix} P_2 \\ -P_4 \end{bmatrix}.$$

We seek a set of $2n$ complementary columns each of which contains a simple unit element but located in a different row position of each of the columns. Clearly

$$\begin{bmatrix} E_1 & P_2 \\ 0_n & -P_4 \end{bmatrix}$$

does not satisfy this criterion, because the jth column of

$$\begin{bmatrix} P_2 \\ -P_4 \end{bmatrix}$$

has its unit element in the ith row position of P_2 corresponding to the same row position as the unit element in the ith column of E_1.

If $i=j$, we merely delete the jth column of

$$\begin{bmatrix} P_2 \\ -P_4 \end{bmatrix}$$

and pick up the required unit element of the jth column of

$$\begin{bmatrix} -P_1 \\ P_3 \end{bmatrix}.$$

This unit will occur in the jth row of P_3. The complementary $2n$-rowed minor now consists of

$$\begin{bmatrix} E_1 \\ 0_n \end{bmatrix} \text{ plus the } j\text{th column of } \begin{bmatrix} -P_1 \\ P_3 \end{bmatrix} \text{ plus } \begin{bmatrix} P_2 \\ -P_4 \end{bmatrix}$$

with its jth column deleted.

If $i\neq j$, the following always yields a $2n$-column complementary minor. Again delete the jth column of

$$\begin{bmatrix} P_2 \\ -P_4 \end{bmatrix} \text{ and substitute the } j\text{th column of } \begin{bmatrix} 0_n \\ E_2 \end{bmatrix}$$

to pick up a unit element in the $n+j$th row position. To maintain complementarity, delete the jth column of

$$\begin{bmatrix} E_1 \\ 0_n \end{bmatrix}$$

and pick up the required unit element in the jth column and jth row of

$$\begin{bmatrix} -P_1 \\ P_3 \end{bmatrix},$$

since its unit element position, with $i\neq j$, has gone undisturbed. The non-zero $2n$-column complementary minor then consists of

$$\begin{bmatrix} E_1 \\ 0_n \end{bmatrix}$$

with its jth column deleted, the jth column of

$$\begin{bmatrix} -P_1 \\ P_3 \end{bmatrix}, \text{ the } j\text{th column of } \begin{bmatrix} 0_n \\ E_2 \end{bmatrix}, \text{ and } \begin{bmatrix} P_2 \\ -P_4 \end{bmatrix}$$

with its jth column deleted. The $[A, B]$ description for P therefore satisfies Theorem 1 and hence P may be realized by use of port-augmenting resistors.

Appendix II

Removal of Boundary Poles from an Immittance Matrix

We show here that if $Z(p)\neq -Z'(-p)$ then $(\overline{Z}(p))^{-1}$ can be found with no boundary poles where:

$$\overline{Z}(p) = Z(p) - \overline{X}(p),$$

and $\overline{X}(p)$ is defined by (38) and is realizable by a lossless n port. The constants $q_i=(A_i, B_i, A, B, a)$ must be chosen to avoid a finite set of forbidden values as given below. The inverse elements of $(\overline{Z})^{-1}$ are given by

$$\bar{y}_{ij}(p) = \frac{A_{ji}(p)}{\bar{\Delta}(p)},$$

as in (39).

1) Just as in the discussion of the one port, choose the numbers q_i so that there is no cancellation of highest powers in the numerators of the $\bar{z}_{ij}(p)$, and if necessary in the determinant $\bar{\Delta}(p)$. This assures that $\bar{\Delta}(p)$ does not exceed the degree of any $A_{ij}(p)$ and hence $\bar{y}_{ij}(p)$ has no poles at ∞.

2) Next we wish to assure the cancellation of finite $j\omega$ boundary poles of A_{ji} by identical boundary factors in $\bar{\Delta}$. To achieve this, observe that in the expansion of $\bar{\Delta}$, each matrix element of $\overline{Z}$ has as coefficient an $(n-1)$-rowed minor. We choose the numbers q_i so that none of these minors vanishes identically, and further so that the boundary zeros of the minors do not coincide with the prescribed poles of $Z(p)$. This means that a finite number of values of the q_i must be avoided and provides that for every boundary pole of $A_{ji}(p)$ due to any element in $\overline{Z}$, a cancelling factor will be present in $\bar{\Delta}$ so the pole will not appear in $y_{ij}(p)$.

3) We must now prevent any boundary zeros from appearing in $\bar{\Delta}$.

If $\bar{\Delta}$ is expanded it will have the form

$$\bar{\Delta} = (\Pi A_i)p^n + (\Pi B_i)\frac{1}{p^n} + f(q_i, p) + \Delta, \qquad (48)$$

where Δ is the determinant of $Z(p)$. Though f will have terms in p^n and $1/p^n$, the coefficients (ΠA_i), (ΠB_i) do not appear again. Thus, whenever Δ vanishes on the boundary, say at $p=j\omega_k$, $[(\Pi A_i)(j\omega_k)^n]+[(\Pi B_i)1/(j\omega_k)^n]$ may be chosen to differ from $f(q_i, j\omega_k)$ and hence Δ will not vanish. Further, at the finite number of boundary points where either the real or imaginary parts of Δ vanish, again the A_i and the B_i may be chosen so that the first two terms of (48) do not combine with Δ and f to produce a boundary zero. Terms in both p^n and $1/p^n$ are included so that no zeros are present at either $p=0$ or $p=\infty$. The only situation under which this technique cannot be carried out is that in which $Z(p)$ is skew-symmetric. Then $\overline{Z}(p)$ is skew-symmetric also, and $\bar{\Delta}$ is purely even or odd (*i.e.*, its real or imaginary parts vanish identically rather than at a finite number of points), so that in general, no choice of the q_i can then avoid a boundary zero. If $Z(p)$ is not skew-symmetric, the addition of the appropriate $\overline{X}$ matrix forces the determinant $\bar{\Delta}$ to be neither even nor odd, even if the elements of $Z(p)$ are themselves purely even or odd. If, therefore, $Z(p)\neq -Z'(-p)$, we may produce a $\overline{Z}(p)$ such that $\overline{Z}^{-1}(p)$ has no boundary poles, and it follows by Theorem 5 that $\overline{Y}(p)$ may be synthesized with only n-positive and n-negative resistors in addition to passive lossless elements. $Z(p)$ is then obtained by adding the realizable n-port reactance structure $\overline{X}(p)$, to the network corresponding to $\overline{Z}(p)$.

BIBLIOGRAPHY

[1] V. Belevitch, "Four dimensional transformations of 4-pole matrices with applications to the synthesis of reactance 4-poles," IRE TRANS. ON CIRCUIT THEORY, vol. CT-3, pp. 105–111; June, 1956.

[2] H. J. Carlin, "General n-port synthesis with negative resistors," PROC. IRE (Correspondence), vol. 48, pp. 1174–1175; June 1960.

[3] H. Aitkin, "Determinants and Matrices," Interscience Publishers, Inc., New York, N. Y.; 1956.

[4] D. C. Youla, "Physical realizability criteria," 1960 IRE CONVENTION RECORD, pt. 2, pp. 181–199.

[5] H. J. Carlin, "Synthesis of Nonreciprocal Networks," *Proc. Symp. on Modern Network Synthesis*, Polytechnic Press of the Polytechnic Inst. of Brooklyn, Brooklyn, N. Y.; 1955.

[6] Y. Oono, "Formal Realizability of Linear Networks," *Proc. Symp. on Active Networks and Feedback Systems*, Polytechnic Press of the Polytechnic Inst. of Brooklyn, Brooklyn, N. Y.; 1960.

[7] V. Belevitch, "On the algebraic structure of formal realizability theory," *Revue H.P.*, vol. IV; 1959.

[8] P. LeCorbeiller, "Matrix Analysis of Electric Networks," Harvard University Press, Cambridge, Mass., John Wiley and Sons, Inc., New York, N. Y., p. 106; 1950.

[9] Y. Oono and K. Yasuura, "Synthesis of finite passive $2n$-terminal networks prescribed scattering matrices," Mem Kyushu Univ. (Eng.), vol. 14, pp. 125–177; 1954.

[10] C. M. Gewertz, "Synthesis of a finite 4-terminal network from its prescribed driving point and transfer functions," *J. Math and Phys.*, vol. 12, pp. 1–257; January, 1932–1933.

Fluctuation Noise in Semiconductor Space-Charge Regions*

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Summary—**In addition to noise arising from the circuit, a semiconductor diodic junction must contain a basic noise source. This basic noise mechanism is due to fluctuations in the ionization state of the impurity atoms which occur at random relative to the probability of ionization, $w = n_{D^+}/n_D$. In an n-type semiconductor, a neutral impurity that suddenly becomes ionized releases a mobile electron whose movement forms a noise current pulse; similarly for the elemental deionization event. The net result is a short-circuited noise current given approximately by the formula,**

$$\overline{I_{\omega_m}{}^2} \approx 2q^2 \frac{n_D A W}{\tau_{D^+}} \frac{(1-w)}{w} \left[-\frac{q(V_B + V_C)}{kT} \right]^{-1} \Delta f.$$

Using typical numbers, this formula indicates that the fundamental noise associated with a 0.3 $\mu\mu$fd depletion layer capacitance at 100 Mc is equivalent to the thermal noise of a 6-ohm series resistor at 40°K and a 0.1-ohm series resistor at 300°K. Experimental verification of this type of noise is not known. When it is observed experimentally, a new method for evaluating some of the basic properties of semiconductors will be available, and computations of minimum noise factor of parametric amplifiers will be possible.

Introduction

SEVERAL writers have examined the noise properties of parametric amplifiers using the nonlinear capacitance associated with the space-charge region of a semiconductor diodic junction. These writers[1–5] have indicated that the significant noise contribution to the performance of the circuit arises from circuit components other than the nonlinear capacitance. These contributions can be largely eliminated by reducing the operating temperature of the circuit and picking appropriate operating frequencies. The question arises: what is the ultimate noise performance that can be expected, and what, if any, noise should be ascribed to the nonlinear capacitor itself? It can be shown on fairly general grounds that a capacitor cannot be the source of noise power.[6,7] However, the capacitor in this case must have certain ideal properties which can only be associated uniquely with a dielectric-free capacitance. As soon as a dielectric is introduced, additional mechanisms are possible which could give rise to noise power. Thus, for the case of the capacitance associated with a semiconductor diodic junction biased in the reverse direction, the capacitance is a monotonically decreasing quantity as a function of increasing bias voltage. This is shown in Fig. 1. A microscopic examination of this capacitance variation would show irregularities as indicated in the insert of Fig. 1. These irregularities might arise from inhomogeneities associated with clusters of impurities. These micro-variations would not generate noise if they are time invariant. However, the impurities have fluctuations in their ionization state, and these fluctuations provide a mechanism whereby energy can be exchanged between the device and its surroundings in much the same manner as fluctuations in a resistor. Thus, the capacitance vs voltage curve could have one shape when it is

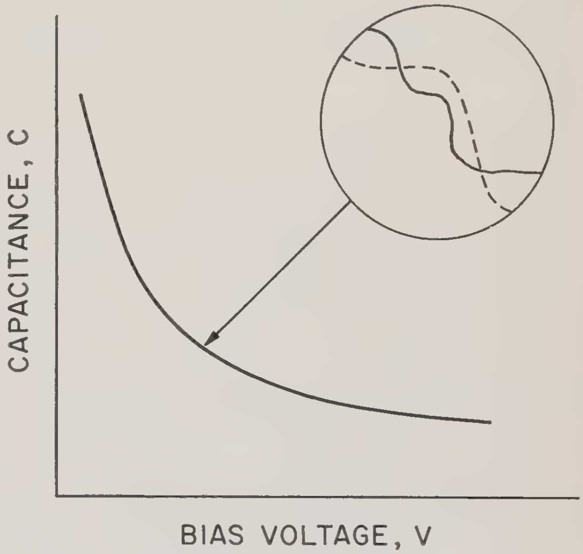

Fig. 1—Typical capacitance variation with bias voltage.

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[1] S. Bloom and K. K. N. Chang, "Theory of parametric amplification using nonlinear reactances," *RCA Rev.*, vol. 18, pp. 578–593; December, 1957.

[2] H. Heffner and G. Wade, "Gain, bandwidth, and noise characteristics of variable-parameter amplifier," *J. Appl. Phys.*, vol. 29, pp. 1321–1331; September, 1958.

[3] M. Uenohara, "Noise consideration of the variable capacitance parametric amplifier," Proc. IRE, vol. 48, pp. 169–178; February, 1960.

[4] R. C. Knechtli and R. D. Weglein, "Low-noise parametric amplifier," Proc. IRE, vol. 48, pp. 1218–1226; July, 1960.

[5] J. C. Greene and E. W. Sard, "Optimum noise and gain-bandwidth performance for a practical one-port parametric amplifier," Proc. IRE, vol. 48, pp. 1583–1590; September, 1960.

[6] F. C. Williams, "Thermal fluctuations in complex networks," *J. IEE*, vol. 81, pp. 751–760; December, 1937.

[7] G. E. Valley, Jr., and H. Wallman, "Vacuum Tube Amplifiers," McGraw-Hill Book Co., Inc., New York, N. Y., p. 506; 1948.

traversed in one direction and a different shape when traversed in the reverse direction. This is shown by the dashed line in the insert of Fig. 1. Under dc bias conditions, the fluctuations in ionization state will be the source of short-circuited mean-square noise current whose value is to be determined.

Equilibrium Probability of Ionization

The probability, w, that an impurity atom is ionized can be determined rather directly, at least for simplified situations,[8] assuming equilibrium conditions to be applicable. Thus, consider a single donor level of concentration, n_D impurities/meter³ with n_{D^+} ionized impurities/meter³, and

$$n_{D^\times} = n_D - n_{D^+} \tag{1}$$

neutral impurities/meter³. If n and p are the electron and hole concentrations, equilibrium space-charge neutrality requires that

$$n = n_{D^+} + p. \tag{2}$$

Furthermore, the laws of mass action state that

$$np = n_i^2, \tag{3}$$

$$n_{D^+}n = K_D n_{D^\times}. \tag{4}$$

These four equations have four unknown quantities, n, p, n_{D^+}, and $n_{D^\times}$ and the following known material constants:

n_i = number of electrons or holes/meter³ in intrinsic material

$$= \sqrt{N_C N_V}\, e^{-E_{CV}/2kT}$$

$$= 25\times 10^{24}\left(\frac{m_n}{m}\right)^{3/4}\left(\frac{m_p}{m}\right)^{3/4}\left(\frac{T}{300}\right)^{3/2} e^{-E_{CV}/2kT};$$

K_D = mass action constant for donor atoms

$$= \frac{N_C}{2} e^{-E_{CD}/kT} \text{ meter}^{-3};$$

N_C = effective density of states in the conduction band

$$= 2\left(\frac{2\pi m_n kT}{h^2}\right)^{3/2}$$

$$= 25\times 10^{24}\left(\frac{m_n}{m}\right)^{3/2}\left(\frac{T}{300}\right)^{3/2} \text{ meter}^{-3};$$

N_V = effective density of states in the valence band

$$= 2\left(\frac{2\pi m_p kT}{h^2}\right)^{3/2}$$

$$= 25\times 10^{24}\left(\frac{m_p}{m}\right)^{3/2}\left(\frac{T}{300}\right)^{3/2} \text{ meter}^{-3};$$

E_{CV} = energy difference between conduction and valence band edges (a positive quantity);

E_{CD} = energy difference between conduction band edge and donor level (a positive quantity);

m_n, m_p = effective masses of electrons in the conduction band and holes in the valence band, respectively;

m = mass of electron;

T = temperature, degrees Kelvin;

h, k = Planck's and Boltzmann's constants, respectively.

The four equations can be combined to give an equation for n_{D+},

$$\left\{1 + \left[1 + 4\left(\frac{n_i}{n_{D^+}}\right)^2\right]^{1/2}\right\}\left(\frac{n_{D^+}}{n_D}\right)^2 + 2\left(\frac{K_D}{n_D}\right)\left(\frac{n_{D^+}}{n_D}\right) - 2\left(\frac{K_D}{n_D}\right) = 0. \tag{5}$$

For moderately doped materials, n_i/n_{D^+} will be a small quantity particularly at temperatures below room temperature, and can therefore be neglected. Then, the solution for the above equation for the fraction of the ionized impurities, n_{D^+}/n_D, which is the probability of ionization, w, is

$$w = \frac{n_{D^+}}{n_D} = \frac{1}{2}\left(\frac{K_D}{n_D}\right)\left\{\left[1 + \frac{4}{\left(\frac{K_D}{n_D}\right)}\right]^{1/2} - 1\right\}. \tag{6}$$

This solution can be approximated in different temperature ranges:

1. At low temperatures where K_D/n_D is small,

$$\frac{n_{D^+}}{n_D} \approx \left(\frac{K_D}{n_D}\right)^{1/2};$$

2) At moderate temperatures where K_D/n_D is not much greater than unity,

$$\frac{n_{D^+}}{n_D} \approx 1 - \left(\frac{n_D}{K_D}\right);$$

(3) At high (room) temperature where K_D/n_D is large,

$$\frac{n_{D^+}}{n_D} \approx 1.$$

[8] See for instance, E. Spenke, "Electronic Semiconductors," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 307–320; 1958.

These approximate solutions together with the exact solution are shown in Fig. 2 where the ionization probability is shown for germanium with a donor impurity concentration, n_D, of 10^{22} atoms/meter^3 corresponding to $\rho \approx 0.2$-ohm cm at 300° K and with the following additional data: $V_{CV}=0.67$ volt; $V_{CD}=0.01$ volt; $m_n/m=0.55$; and $m_p/m=0.37$.

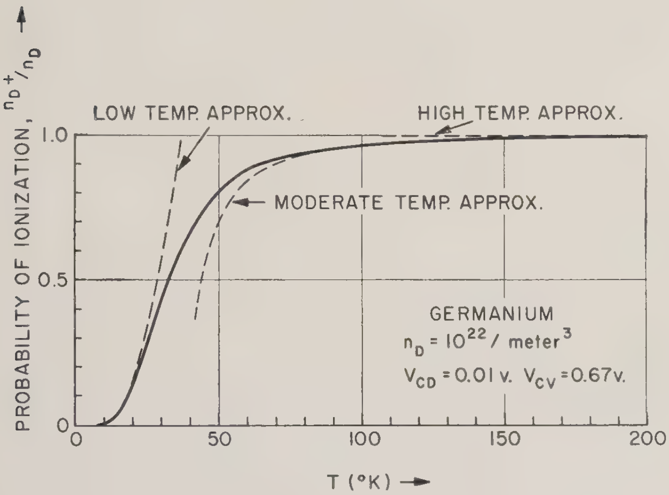

Fig. 2—Equilibrium probability of ionization as a function of temperature.

Nonequilibrium Probability of Ionization

The probability of ionization of a single donor level using a spin degeneracy factor of $\frac{1}{2}$ corresponding to a donor atom with a single valence s—electron is[8]

$$w = \frac{n_{D^+}}{n_D} = \frac{1}{1 + \frac{N_C}{K_D} e^{-(E_{CF}/kT)}}. \tag{7}$$

Here, E_{CF} is the energy difference between the conduction band edge and the Fermi level (usually a positive quantity) defined by

$$E_{CF} = -\zeta\left(\frac{n}{N_C}\right) \approx -kT \ln\left(\frac{n}{N_C}\right). \tag{8}$$

The approximate expression above for E_{CF} is applicable when $n/N_C \ll 1$, in which case (7) becomes

$$w = \frac{n_{D^+}}{n_D} \approx \frac{1}{1 + \frac{n}{K_D}}. \tag{9}$$

This result can also be obtained by substituting (1) into (4). If a further assumption of $n \approx n_{D^+}$ is employed, (5) (with $n_i = 0$) would result. However, within the depletion region $n \neq n_{D^+}$, but rather

$$n(x) \approx n(W) e^{+qV(x)/kT}, \tag{10}$$

where $V(x)$ is the voltage (negative in value) of x including a small contact voltage measured with respect to zero potential deep within the n-type semiconductor where the equilibrium electron density is $n(W)$. Thus, the probability of ionization would vary from the equilibrium value given by (6) at the edge of the depletion region, $x=W$, to essentially unity at the biased junction, $x=0$, where $n \ll K_D$.

Noise Effects

As the operating temperature of the diode is reduced, the probability that an atom is ionized becomes less, and the capacitance at a fixed bias will decrease as the noise fluctuation arising from it increases. Thus, if a parametric amplifier using a semiconductor nonlinear capacitor is refrigerated to reduce circuit noise, the fluctuation noise associated with the capacitor itself will increase. However, the refrigeration temperature could be reduced rather considerably (say 30°–40°K near liquid hydrogen temperature) before fluctuation capacitor noise might become appreciable. It appears likely then that the circuit noise factor might have a minimum value at some lower temperature. The fluctuation phenomena considered here is similar to the fluctuation phenomena associated with quantum transitions in crystals used in maser amplifiers,[9–11] and an interesting question arises as to which of the two devices have lower fundamental noise potentialities.[12] It is interesting to note that maser-type amplifiers must be refrigerated to reduce noise arising from spontaneous (in contrast with induced) energy transitions. In direct contrast, parametric amplifiers employing nonlinear semiconductor capacitances will have greater inherent noise at low temperatures in comparison to room temperature.

In order to minimize the fluctuation noise generation associated with a reverse-biased junction, it is advisable to use impurities with very small ionization potentials. Also, impurities with large ionization potentials must be rigorously eliminated. It appears likely that when devices are made which approach the fundamental noise condition discussed here, the residual noise can be used as a means for evaluating impurity levels in semiconductors and as a means of critical evaluation in the same manner that noise is used as evaluation of electron devices.

[9] J. P. Wittke, "Molecular amplification and generation of microwaves," Proc. IRE, vol. 45, pp. 291–316; March, 1957.

[10] M. W. Muller, "Noise in molecular amplifier," *Phys. Rev.*, vol. 106, pp. 8–12; April 1, 1957.

[11] K. Shimoda, H. Takahasi, and C. H. Townes, "Fluctuations in amplification of quanta with application of maser amplifiers," *J. Phys. Soc.*, (Japan) vol. 12, pp. 686–700; June, 1957.

[12] M. W. P. Strandberg, "Inherent noise of quantum-mechanical amplifiers," *Phys. Rev.*, vol. 106, pp. 617–620; May 15, 1957.

Elemental Event Analysis

Consider a depletion region extending from $x'=0$ at the diodic junction where the bias voltage plus internal contact voltage is (V_B+V_C) into the semiconductor to the limit of the depletion at $x'=W$ where the voltage is zero. Assume that at the plane $x'=x$, a neutral donor atom becomes ionized at $t=0$. The mobile electron released by this event moves towards $x'=W$ and gives rise to a current. The instantaneous magnitude of this current can be evaluated by considering that the charge moves through an incremental voltage, $dV=-Edx'$, and the resulting energy change can be related to an external movement of charge, $dQ=idt$. Thus

$$i = q \frac{E}{(V_B - V_C)} \frac{dx'}{dt}. \tag{11}$$

For small electric fields, E, the mobility is of the electron, μ_n, independent of E, and $dx'/dr=-\mu_n E$. If further the values of $E=-(qn_D{}^+/K\epsilon_o\ (W-x')$ and $(V_B+V_C) = -(qn_D{}^+/K\epsilon_o)\ (W^2/2)$ for the depleted semiconductor are used in (11),

$$i = \frac{2q}{\tau_r}\left(\frac{W - x'}{W}\right)^2 \tag{12}$$

is obtained where τ_r is the relaxation time of electrons in the n-type semiconductor, $\tau_r=\rho K\epsilon_o$. The subsequent position of the released electron moving towards $x'=W$ is given by

$$x' = W - (W - x)e^{-t/\tau_r}\ (x < x' < W), \tag{13}$$

and the elemental current associated with this event is

$$i(t) = \frac{2q}{\tau_r}\left(\frac{W - x}{W}\right)^2 e^{-2t/\tau_r}. \tag{14}$$

The transit time, t_t, of an electron moving through the depleted region can be obtained by using (13) with $x=0$ and $x'=W-x_T$. The result is

$$t_t = \tau_r \ln\left(\frac{W}{x_T}\right) = \frac{\tau_r}{2} \ln\left[-\frac{q(V_B + V_C)}{kT}\right]. \tag{15}$$

Here x_T is a distance corresponding to a variation of kT/q volts at $x=W$ given by

$$x_T = \left[\frac{2K\epsilon_o kT}{n_D{}^+ q^2}\right]^{1/2}. \tag{16}$$

The evaluation is carried out to a distance x_T short of the depletion width to avoid an infinite transit time that would otherwise result.

In addition to ionization events there are also deionization events present which give rise to current. While deionization can occur in a variety of ways, the event that is considered most probable is that the ionized donor atom at $x'=x$ captures an electron from a nearby electron-hole pair and the released mobile hole moves towards $x'=0$ with the subsequent x' position given by

$$x' = W\left(1 - e^{t - t_t'/b\tau_r}\right), \qquad (0 < x' < x). \tag{17}$$

The elemental current associated with this event is

$$i'(t) = \frac{2q}{b\tau_r}\, e \,\frac{2(t - t_t')}{b\tau_r}. \tag{18}$$

A typical ionization event occurring at $x'=x_0$ would give rise to an exponential current pulse as shown in Fig. 3(a); a typical deionization event gives rise to an exponentially rising current pulse as shown in Fig. 3(b). Other typical events also occur, but these are considered sufficiently improbable to be of negligible importance. The current pulses of Fig. 3 therefore constitute elemental events of the noise phenomena. Since the current pulses have been formulated on the basis that the voltage distribution in the semiconductor remains unchanged throughout the duration of the pulse, an incremental short circuit at the terminals of the capacitor is implied. This is a simplification of the analysis for the same reason that noise properties of a resistor are generally formulated on the basis of the mean-square short-circuited current rather than the mean-square open-circuited voltage.

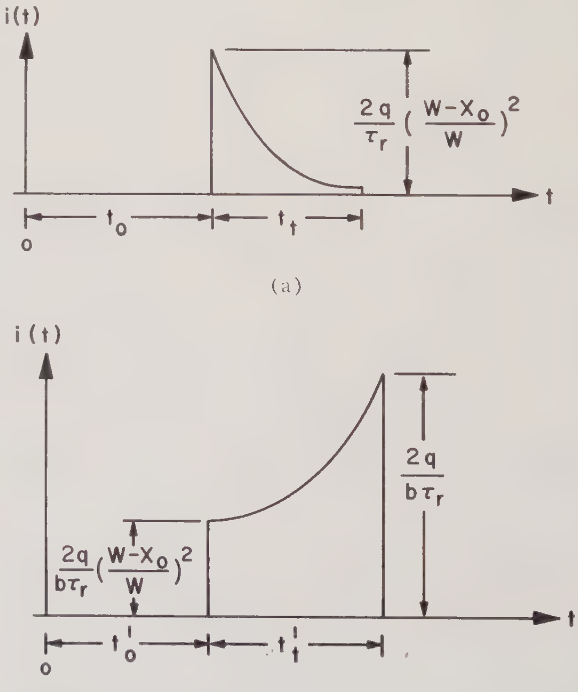

Fig. 3—Current pulses: (a) as arising from a typical ionization fluctuation occurring at x_0; (b) as arising from a typical deionization fluctuation occurring at x_0.

A very useful noise theorem[13] can now be used to obtain the mean-square short-circuited noise current, $I_{\omega_n}^2$, from the elemental current pulses and statistics associated therewith. Thus,

$$\overline{I_{\omega_n}^2} = 2\frac{M}{T}\overline{|F(\omega_n)|^2}\Delta f \tag{19}$$

where M/T = pulse events/second averaged over a time interval, T, long compared to the duration of the pulse, with T establishing the frequency interval, $\Delta f = 1/T$ so $\omega_n = 2\Pi n/T$.

$$F(\omega_n) = \int_0^{t_t} i(t)e^{-j\omega_n t}dt$$

is the complex Fourier analysis of an elemental current pulse starting at $t=0$. The mean-square short-circuited noise current will be obtained by adding together the contributions from ionization and deionization events occurring in an elemental volume $A\Delta x$ and summing these contributions from $x=0$ to $x=W$. The impurity atoms are assumed to change their state independently and to make their current contribution separately. Thus, only one electron or hole is assumed to be moving at a given time. This assumption appears justified only because the carrier transit time, t_t, is usually much smaller than the lifetime of the neutral atom, $\tau_D^\times$, or the lifetime of the ionized atom, τ_D^+. Multiple events that do occur will add significantly to the mean-square noise current so the results to be derived will represent a lower limit to the actual noise.

Fluctuation Analysis

Over a time interval, T, M independent ionizations occur associated within an elemental volume, $A\Delta x$, and

$$\frac{M}{T} = \frac{\text{number of ionization events}}{\text{(unit time)}} = \frac{n_D^\times}{\tau_D^\times}A\Delta x. \tag{19}$$

The mean-square noise current associated with the M independent ionization events in a period T is

$$\overline{\Delta I_{\omega_n}^2} = \frac{2q^2 n_D^\times}{\tau_D^\times}\left(\frac{W-x}{W}\right)^4 |F_x(\omega_n)|^2\Delta f A\Delta x \tag{20}$$

where

$$F_x(\omega_n) = \frac{1-e^{-(2/\tau_r+j\omega_n)t_t}}{\left(1+j\frac{\omega_n\tau_r}{2}\right)}. \tag{21}$$

M' independent deionization events occur over a period, T, and

$$\frac{M'}{T} = \frac{\text{number of deionization events}}{\text{(unit time)}} = \frac{n_D^+}{\tau_D^+}A\Delta x. \tag{22}$$

The mean-square noise current associated with the M' independent deionization events in a period T is

$$\overline{\Delta I_{\omega_n}^2} = \frac{2q^2 n_D^+}{\tau_D^+}|F_x'(\omega_n)|^2\Delta f A\Delta x \tag{23}$$

where

$$F_x'(\omega_n) = -\left[\frac{e^{-2t_t'/b\tau_r} - e^{-j\omega_n t_t'}}{1-\frac{j\omega_n b\tau_r}{2}}\right]. \tag{24}$$

Since on the average, local static conditions prevail, $n_D^\times/\tau_D^\times = n_D^+/\tau_D^+$, and the total short-circuited noise current is

$$\overline{I_{\omega_n}^2} = \frac{2q^2}{\tau_D^\times}A\Delta f\int_0^W n_D^\times\left[\left(1-\frac{x}{W}\right)^4|F_x(\omega_n)|^2 + |F_x'(\omega_n)|^2\right]dx, \tag{25}$$

where use has been made of the fact that $\tau_D^\times$ is independent of x. The approximate dependence of $n_D^\times$ upon x can be obtained by using (9), (10), and the parabolic voltage variation through the semiconductor, $V(x) = (V_B+V_C)\,(1-x/W)^2$. Thus

$$\frac{n_D^\times}{n_D} = 1 - \frac{n_D^+}{n_D}\frac{n/K_D}{1+n/K_D} \approx \frac{\frac{n(W)}{K_D}\exp\left\{+\frac{q}{kT}(V_B+V_C)\left(1-\frac{x}{W}\right)^2\right\}}{1+\frac{n(W)}{K_D}\exp\left\{+\frac{q}{kT}(V_B+V_C)\left(1-\frac{x}{W}\right)^2\right\}}. \tag{26}$$

Since n_D^+/n_D quickly approaches unity as x becomes smaller than W, it is apparent that the bulk of the noise will arise from the semiconductor near the edge of the depletion region. Thus, the effect of the voltage through the space-charge region is to suppress noise generation, and following the practice of North,[14] a space-charge noise suppression factor, Γ^2, can be introduced with (25) being written as

$$I_{\omega_n}^2 = \Gamma^2 2q^2\frac{n_D AW(1-w)}{\tau_D^\times}\Delta f \tag{27}$$

[13] See for instance, S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. J.*, vol. 23, pp. 282–332; July, 1944; or, A. van der Ziel, "Noise," Prentice-Hall, Inc., New York, N. Y., pp. 321–323, 1954.

[14] B. J. Thompson, D. O. North, and W. A. Harris, "Fluctuations in space-charge-limited currents at moderately high frequencies," *RCA Rev.*, vol. 4, pp. 441–472; April, 1940.

with w given by (6) and where

$$\Gamma^2 = \frac{1}{n_D(1-w)W}\int_0^W n_D{}^{\times}\left[\left(1-\frac{x}{W}\right)^4 |F_x(\omega_n)|^2 + |F_x'(\omega_n)|^2\right]dx. \quad (28)$$

Values of Γ^2 could be obtained by carrying out the indicated integration by graphical or other numerical means.

An approximate expression for Γ^2 can be obtained by recognizing that in the vicinity of $x=W$ where $n_D{}^{\times}$ is significant $(1-x/W)|F_x(\omega_n)|^2\approx 0$ (ionization events can be neglected since velocity of electron is small) and $|F_x'(\omega_n)|^2\approx 1$ as long as the frequency range of interest is restricted to low enough frequencies so that $\omega_n\tau_r \ll 1$. Then

$$\Gamma^2 \approx \frac{1}{(1-w)}\int_0^1 \frac{\dfrac{n(w)}{K_D}\exp\left\{\dfrac{q(V_B+V_C)}{kT}\alpha\right\}}{1+\dfrac{n(W)}{K_D}\exp\left\{\dfrac{q(V_B+V_C)}{kT}\alpha\right\}}\,d\alpha. \quad (29)$$

Note that $\Gamma^2\to 1$ as $W\to 0$ corresponding to small values of (V_B+V_C). Over much of the range of practical interest

$$\frac{n(W)}{K_D} = \frac{1-w}{w} < 1,$$

and the denominator in the above expression for Γ^2 can be adequately approximated by unity with the result that

$$\Gamma^2 \approx \frac{1}{w}\left[\frac{1-\exp\left(\dfrac{q(V_B+V_C)}{kT}\right)}{-\dfrac{q(V_B+V_C)}{kT}}\right] \approx \frac{1}{w}\left[-\frac{q(V_B+V_C)}{kT}\right]^{-1}. \quad (30)$$

The last approximation for Γ^2 is valid provided

$$-\frac{q(V_B+V_C)}{kT}$$

is large compared to unity. Subject to the indicated approximations, (27) becomes

$$\overline{I_{\omega_n}{}^2} \approx 2q^2\,\frac{n_D A W}{\tau_D{}^{\times}}\,\frac{(1-w)}{w}\left[-\frac{q(V_B+V_C)}{kT}\right]^{-1}\Delta f, \quad (31)$$

TABLE I
CALCULATIONS OF NOISE

Quantity	Units	$T=40°$K	$T=300°$K	Remarks
Bias, (V_B+V_C)	volts	-1	-1	
Area, A	cm^2	10^{-5}	10^{-5}	
Impurities, n_D	cm^{-3}	10^{16}	10^{16}	
Probability, w		0.67	0.9971	See Fig. 2
Depletion, W	microns	0.515	0.420	
Capacitance, C	$\mu\mu$fd	0.275	0.330	
Resistivity, ρ	ohm cm	0.0937	0.187	
Relaxation time, τ_r	$\mu\mu$sec	0.0123	0.246	
Transit time, t_t	$\mu\mu$sec	0.0348	0.449	Eq. (15)
Ionized lifetime, $\tau_D{}^{\times}$	μsec	2.50	95	Ref. 18
Neutral lifetime, $\tau_D{}^{\times}$	μsec	1.23	0.276	
Noise, $\overline{I_{\omega n}{}^2}$	$(\mu\mu\text{amps})^2$	0.366 $\times 10^{-3}\,\Delta f$	0.059 $\times 10^{-3}\,\Delta f$	Eq. (31)
Equivalent noise conductance	μmhos	0.166	0.00355	$4kTG\,\Delta f$
Noise, $\overline{V_{\omega n}{}^2}$	$(\mu\mu\text{volts})^2$	12.4 $\times 10^3\,\Delta f$	1.99 $\times 10^3\,\Delta f$	at $f=100$ Mc
Equivalent noise resistance	ohms	5.60	0.12	$4kTR\,\Delta f$
Γ^2		5.14 $\times 10^{-3}$	26×10^{-3}	Eq. (30)

with w being the equilibrium probability of ionization as given by (6). This formula was used to compute noise properties of a device at two temperatures, 40°K and 300°K, as shown in Table I. It is seen that the space-charge noise suppression factor has a significant effect in reducing the noise. The biggest uncertainty in the calculations is an appropriate value for $\tau_D{}^{x}$. Low temperature measurements of $\tau_D{}^{+}$ have been reported by Wertheim[15] for indium, and by Koenig[16] for antimony; Lax[17] has examined the problem theoretically. The values of $\tau_D{}^{+}$ shown in Table I are obtained by temperature extrapolation using observed $T^{1.8}$ dependency of experimental data by Ascarelli[18] for antimony doped germanium.

[15] G. K. Wertherim, "Temperature dependence of capture cross sections of indium in silicon," *Bull. Am. Phys. Soc.*, series II, vol. 2, p. 314; September 5, 1957.

[16] S. Koenig, "Recombination of thermal electrons in n-type germanium below 10°K," *Phys. Rev.*, vol. 110, pp. 988–990; May 15, 1958.

[17] M. Lax, "Giant Traps," *J. Phys. Chem. Solids*, vol. 8, pp. 66–73; January, 1959.

[18] G. Ascarelli and S. C. Brown, "Recombination of electrons with donors in n-type germanium," *Phys. Rev.*, vol. 120, pp. 1615–1626; December 1, 1960.

Noise Contributions vs Temperature

Some appreciation of the noise contributions of a variable capacitance device can be obtained from computations made as a function of temperature. The calculations were carried out for the same diode device considered before. The results are tabulated in Table I and are shown in Fig. 5. R_{eq} in Fig. 5 is the resistance whose Nyquist noise at the operating temperature would be equivalent to the diode noise. It is evaluated at 100 Mc using a transformation of the noise current generator to a noise voltage generator as shown in Fig. 4. $\overline{I_{\omega_n}^2}$ is computed using w as shown in Fig. 2, and $\tau_D^\times$ is determined as mentioned above. There is of course considerable uncertainty concerning the vaidity of extrapolating Ascarelli's measurements[18] made near 4°K to 300°K.

As temperature is reduced, C will diminish as $w^{1/2}$ and is plotted in Fig. 5. Diodes invariably have a small series resistance, R_S, which is a significant source of device noise, and its variation with temperature is also shown in Fig. 5. R_S is determined by arbitrarily choosing a value of 4.74 ohms at 300°K to correspond to a cutoff frequency of 100 kMc; and assuming that the resistance is proportional to the bulk resistivity; the temperature variation is introduced via the variation in mobility and the number of ionized impurities. The temperature variation of mobility is calculated by using the best advice as contained in the survey papers by Conwell,[19] and the results are similar to measured conductivity variations that she attributes to Debye. The plot of Fig. 5 clearly indicates how R_S diminishes initially as temperature is reduced and R_{eq} gradually increases. The two resistances are equal at 60°K, and the 100-Mc noise for this particular diode therefore goes through a broad minimum near this temperature. At higher operating frequencies R_{eq} is smaller. The $R_{eq} = R_S$ point would occur at a lower temperature, and the noise minimum temperature may correspond more nearly to the minimum R_S temperature.

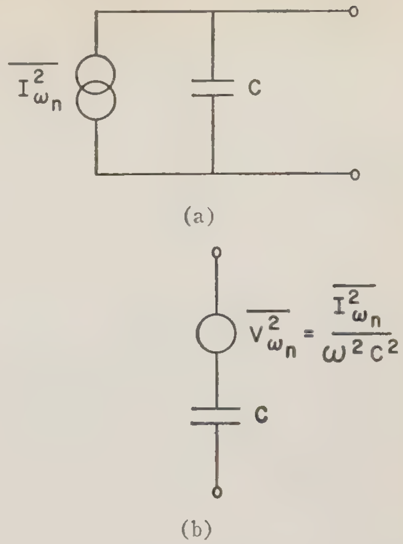

Fig. 4—Representation of barrier capacitance with (a) mean-square noise current generator; (b) mean-square noise voltage generator.

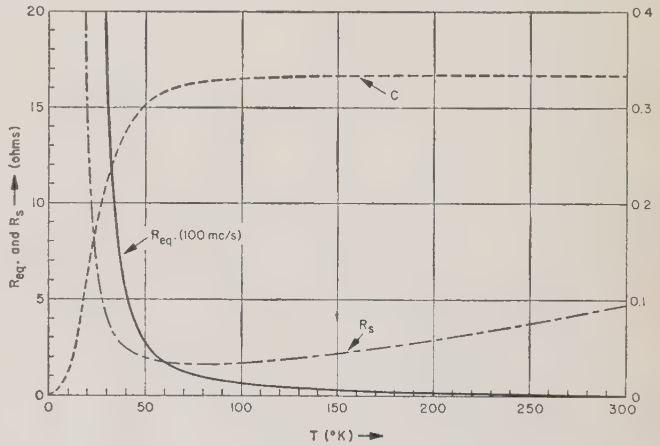

Fig. 5—Temperature variation of equivalent noise resistance, R_{eq}, and of series resistance, R_S, for a germanium diode of capacitance, C, at a fixed bias of 1 volt.

Conclusions

The analytical evaluation of noise inherent to the nonlinear capacitance should provide a stimulus for the experimental verification of its presence. With care, experimental evaluation should be possible. Improvement of measurement techniques will then lead to possible further reduction in the parametric device noise and may also provide a method for evaluating material properties such as impurity energy levels and lifetimes of ionized impurities. The analytical formulation of the inherent noise should permit the determination of the basic noise performance of parametric diodes in various circuits and should provide a good basis for comparing the limiting noise capability of these diodes with other competitive devices such as masers, electron beams using space-charge waves, ferromagnetic modulators, Esaki diodes,[20] etc.

The noise development given here can be extended as required to include effects arising when more then one impurity or neutralized impurities are present, when correlation effects are important due to multiple simultaneous ionization or deionization events, and when multiple events occur during a transit time. Correlation effects which are assumed negligible in this analysis may well be significant since the noise output is rather sensitive to the presence of multiple events.

[19] E. M. Conwell, "Properties of silicon and germanium," Proc. IRE, vol. 40, pp. 1327–1337, November, 1952; also, E. M. Conwell, "Properties of silicon and germanium: II," Proc. IRE, vol. 46, pp. 1281–1300, June, 1958.

[20] J. J. Tiemann, "Shot noise in tunnel diode amplifiers," Proc. IRE, vol. 48, pp. 1418–1423; August, 1960.

Electrodeless Measurement of Semiconductor Resistivity at Microwave Frequencies*

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Summary—**A new microwave technique for the measurement of conductivity of semiconductors has been explored and provides agreement with more conventional methods. The proposed technique depends upon the absorption of the microwave power being propagated through a semiconductor medium. This eliminates the need for electrode attachment, making the experimental aspects of the measurement more simple. In addition, since the microwave method depends more on bulk properties, it may be less subject to error due to surface leakage or crystal imperfections in the semiconductor.**

Introduction

During the past few years there has been an increased interest in the interactions of microwave energy and semiconductors. In particular, microwave techniques have been used to determine fundamental properties of such materials as germanium and silicon.[1] Recently an electrodeless method of measuring lifetime was described in which the changes in microwave absorption, due to the injection of excess minority carriers and their subsequent decay, was utilized as the indicating device.[2]

In this paper an electrodeless method of measuring resistivity of semiconductors is discussed. Single crystal samples of different resistivities are placed in a waveguide and, by determining the transmission of power through the sample, a direct determination of the resistivity can be obtained.

This new proposed method has several advantages. First, since it represents an arrangement almost identical with the "lifetime apparatus," both resistivity and lifetime can be measured in the same equipment. Second, in the proposed technique for resistivity measurements, the problem of contacts is eliminated. The construction of "ohmic contacts" on some of the newly utilized semiconductors such as high-purity silicon, gallium arsenide and gallium phosphide is a difficult problem, particularly at low temperatures where contacts which have been considered as ohmic have often been found to rectify, inject and/or extract minority carriers. The third advantage of the new technique is that surface states or bulk nonuniformities such as grain boundaries will obscure direct-current measuring methods but they are generally not as important when using microwave absorption. This is due to the fact that for high-resistivity materials, the amount of microwave energy absorbed at the surface is negligible. All of the absorption of power can be looked upon as occurring in the bulk volume. One disadvantage of the microwave technique is that the material must be carefully shaped; for very high accuracy, the sides of the sample must be parallel and fitted into the waveguide with a minimum of air space between the guide walls and the sample.

Description of Method

Consider a transmission system consisting of a waveguide in which electromagnetic energy is being propagated in the TE_{10} mode. Assume that a slab of semiconductor is inserted in the waveguide as shown in Fig. 1,

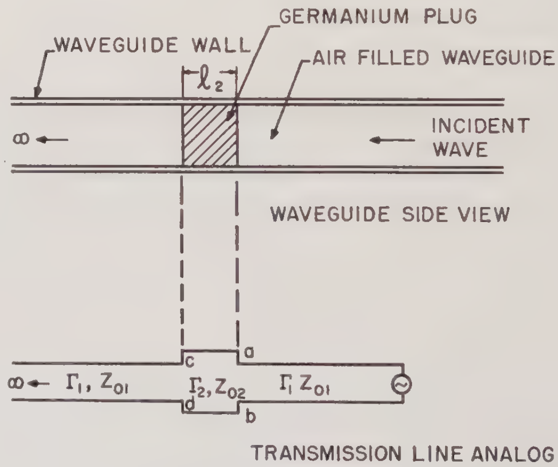

Fig. 1—Equivalent circuit for a slab of semiconductor in a waveguide.

where the large surface of the semiconductor slab is oriented perpendicularly to the direction of propagation. Here the electromagnetic wave will exhibit multiple internal reflections and will be partly transmitted, reflected and absorbed. Since it is relatively easy to measure transmitted power, the following theory will apply to the measurement of E_o/E_{in}, or the ratio of transmitted electric field to the incident electric field.

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[1] A. F. Gibson, "Progress in Semiconductors," John Wiley & Sons, Inc., New York, N. Y., vol. 2, pp. 229–245; 1957.

[2] A. P. Ramsa, H. Jacobs, and F. A. Brand, "Microwave techniques in measurement of lifetime in germanium," 1959 IRE National Convention Record, pt. 3, pp. 159–168. "Further consideration of bulk lifetime measurement with a microwave electrodeless technique," Proc. IRE, vol. 48, pp. 229–233, February, 1960.

This ratio is given by (1) and is derived in the Appendix. Here,

$$\frac{E_o}{E_{in}} = r_t\left(\cosh \Gamma_2 l_2 - \frac{Z_{02}}{Z_{ab}} \sinh \Gamma_2 l_2\right), \tag{1}$$

where

$$r_t = \frac{2Z_{ab}}{Z_{ab} + Z_{01}}, \tag{2}$$

and

$$Z_{ab} = Z_{02}\left(\frac{Z_{01} + Z_{02} \tanh \Gamma_2 l_2}{Z_{02} + Z_{01} \tanh \Gamma_2 l_2}\right). \tag{3}$$

In (1)–(3),

Z_{01} = the impedance of the line (waveguide) in air and is equal to Z_{03}.

Z_{02} = the impedance of the waveguide filled with germanium and is a function of conductivity and wavelength.

Z_{ab} = the impedance of the germanium at the front surface.

Γ_1 = the propagation constant in air in the waveguide.

Γ_2 = the propagation constant in germanium in the waveguide.

l_2 = the thickness of the slab of semiconductor.

Each of these terms is computed, as indicated in the Appendix, as a function of conductivity at the frequency of 9.549 kMc in X-band waveguide.

Now if the thickness of the semiconductor plug is held constant (*i.e.*, 0.392×10^{-2} meters) and the conductivity varied, using (1)–(3) one can obtain a calculation of the magnitude of the ratio of transmitted electric field to the incident field, $|E_o/E_{in}|$, as a function of conductivity. The results of these calculations are shown in Fig. 2. In the use of these data one can convert the field ratio to power ratio by squaring $|E_o/E_{in}|$.

Next we shall describe experiments where, by measuring the transmitted power through the germanium plug, one can use Fig. 2 to determine the conductivity of the specimen. In these experiments the conductivity, σ, could be varied by heating the semiconductor sample. Conductivity could then be compared with temperature, and the energy band gap of the material could be calculated.[3] At the same time, the resistivity of intrinsic germanium at 300°K could be checked.

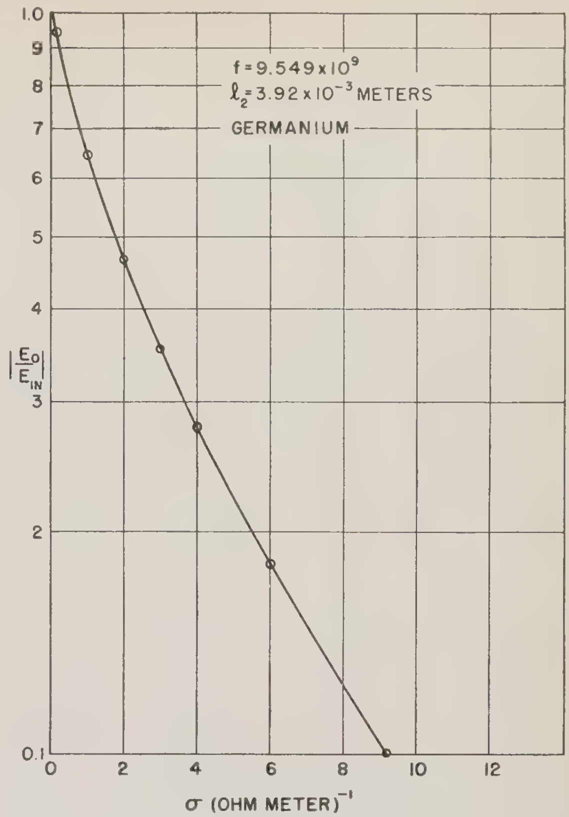

Fig. 2—The computed magnitude of E_o/E_{in} as a function of conductivity for germanium. The frequency is 9.549×10^9 cps, the thickness of the slab is 3.92×10^{-3} meters, the waveguide cutoff wavelength was taken as 4.56×10^{-2} meters.

More specifically, the germanium samples were cut and polished to fit an X-band thermal control unit as shown in Fig. 3. A snug fit could be obtained with respect to the width and height. The sample length was held fixed at 0.392×10^{-2} meters, which is approximately one-half wavelength for 9.549 kMc radiation in germanium. Next, the system was matched with the thermal control unit empty and the VSWR maintained $\leqq 1.03$. The sample was then inserted in the microwave system shown in Fig. 4. The ratio of $|P_o/P_{in}|$ was measured and it was established that varying the exact position of the sample with respect to the distance of the load or generator would cause no change in $|P_o/P_{in}|$. The values of $|E_o/E_{in}|$ were then determined as a function of the temperature of the semiconductor. Using Fig. 2, σ could be determined as a function of the reciprocal of temperature. Results are indicated in Figs. 5 and 6. These data show agreement with Conwell[4] who relates the conductivity of intrinsic germanium to temperature using conventional dc resistivity measuring techniques, such as dc current methods or "four-point probe" techniques.

[3] Here the relation $\sigma = \sigma_0 e^{-E_g/2kT}$ was employed, where σ is the conductivity, k the Boltzmann's constant, T is the temperature in degrees Kelvin, and E_g the band gap.

[4] E. M. Conwell, "Properties of silicon and germanium," Proc. IRE, vol. 40, pp. 1327–1337; November, 1952.

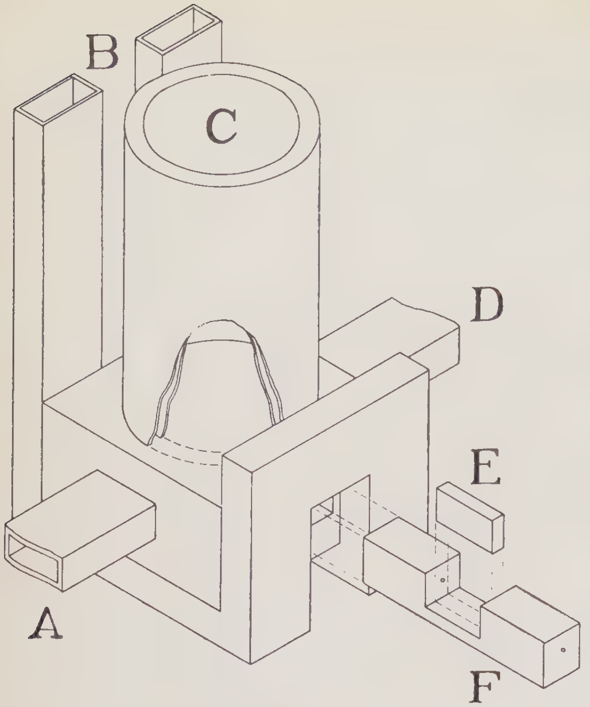

Fig. 3—The thermal control unit. The entrance A designates the input and D the output. Both B and C designate openings for the fluid used to control the temperature of the large metal block. The holder F can slide into position in the waveguide after the sample E is inserted. Openings were drilled in F for probes, if desired. The transmission system is for X-band frequencies.

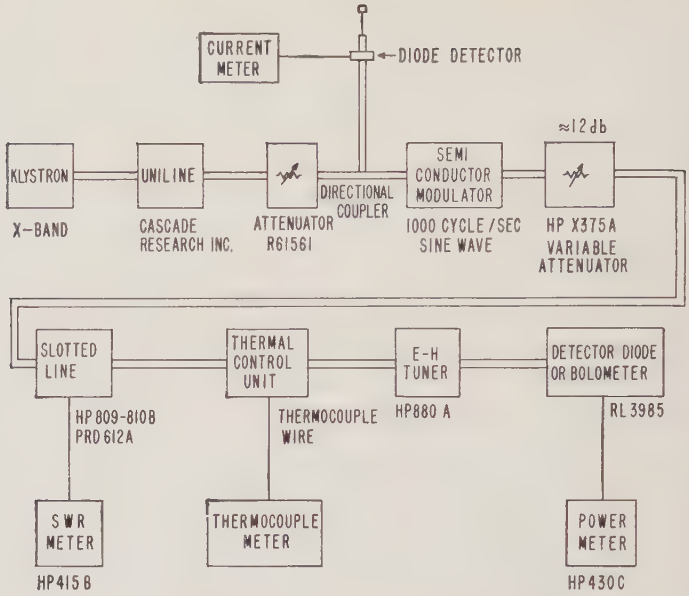

Fig. 4—Schematic diagram for microwave experimentation.

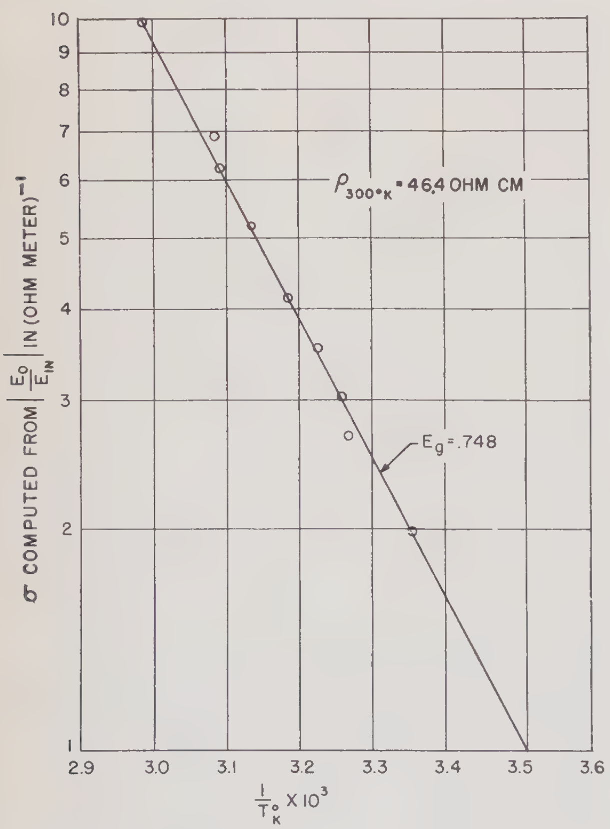

Fig. 5—The conductivity σ as a function of the reciprocal of temperature. The σ is given in mks, temperature in °K. The conductivity was taken from experimental values of $|E_o/E_{in}|$ by means of the computed data shown in Fig. 2.

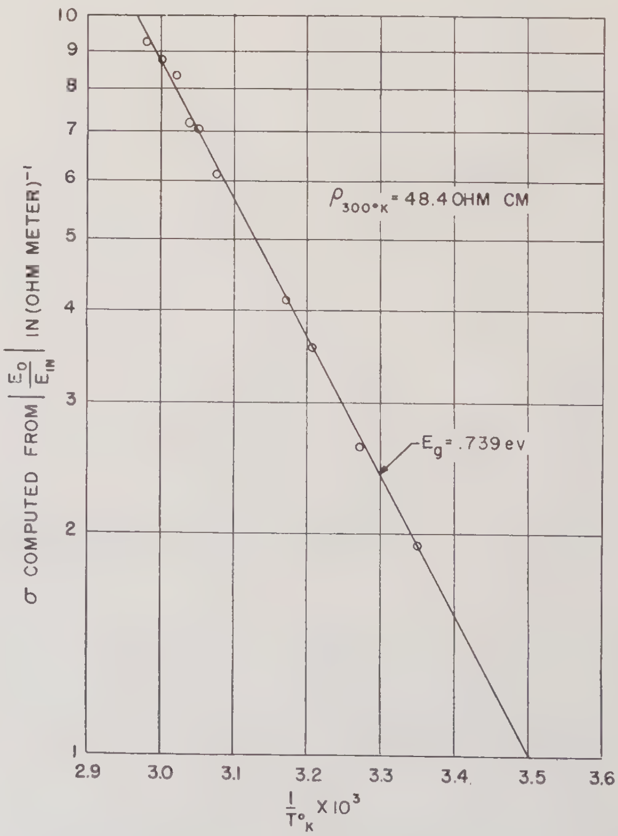

Fig. 6—The conductivity σ as a function of the reciprocal of temperature. The σ is given in mks, temperature in °K. The conductivity was taken from experimental values of $|E_o/E_{in}|$ by means of the computed data shown in Fig. 2.

In Table I we see the results of the measurements indicating the values of every band gap E_g and resistivity ρ in ohm-cm, as determined by these experiments together with values found in the literature.[4]

Experiments of a similar nature have been started with silicon but have not been extensively studied. Initial results, however, have been encouraging. In one sample, for instance, the values of $|E_o/E_{in}|$ vs σ were calculated using $\epsilon = 12 \times 8.854 \times 10^{-12}$ and $l_2 = 0.474 \times 10^{-2}$ meters. The calculated data are shown in Fig. 7. Next, using the same technique of measuring $|E_o/E_{in}|$, the value of σ was determined at room temperature. The microwave method indicated $\rho \cong 5000$ ohm-cm. The "four-probe method" indicated $\rho = 3000$ to 6000 ohm-cm. This again represents fairly good agreement.

TABLE I
ENERGY BAND GAP AND RESISTIVITY OF GERMANIUM

Trial Run	Experimental Band Gap Near 300°K in ev	Experimental Resistivity in ohm-cm at 300°K	Band Gap Near 300°K as shown in Literature[4]	Resistivity in ohm-cm at 300°K as shown in Literature[4]
1	0.70	46.1	0.72	47.0
2	0.75	45.3	0.72	47.0
3	0.75	52.5	0.72	47.0
4	0.75	46.4	0.72	47.0
5	0.74	48.4	0.72	47.0
Average Value	0.74	47.7	0.72	47.0

CONCLUSION

The microwave technique for measurement of conductivity of semiconductors has been demonstrated and appears to give good agreement with expected values obtained by other methods. The new technique has advantages particularly where ohmic contacts may be a problem or where the "four-point probe" methods may be difficult. Although these experiments were run at 9.546 kMc, other microwave frequencies are possible. This would permit a greater range of size of sample. For instance, if a new material to be studied were not available in larger size crystals, a higher frequency would be more applicable. The 10-kMc region appears, however, most generally useful for silicon and germanium. Further computations for germanium are shown in Fig. 8 where $f = 10^{10}$ cps, a relatively convenient frequency, and $|E_o/E_{in}|$ is shown as a function of length and conductivity in standard X-band waveguides (4.56×10^{-2} meters width). For a given sample length, the $|E_o/E_{in}|$ vs σ data can be taken directly from these plots. The phase angle of E_o/E_{in} has also been computed. Although not directly of interest in these experiments, it can be pointed out that the magnitude of reflection, the reflection-phase angle or the phase angle of E_o/E_{in}, can all be

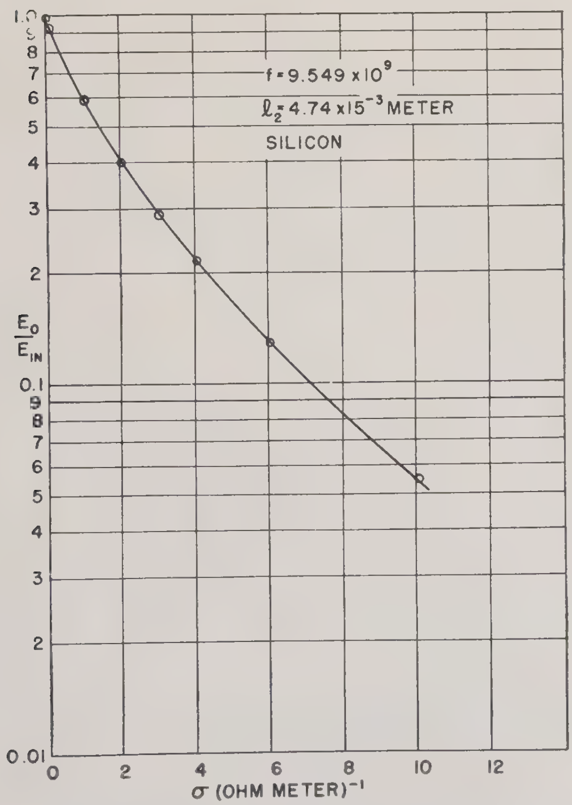

Fig. 7—The computed magnitude of E_o/E_{in} as a function of conductivity for silicon. The frequency is 9.549×10^9 cps, the thickness of the slab is 4.74×10^{-3} meters, the waveguide cutoff wavelength was taken as 4.56×10^{-2} meters.

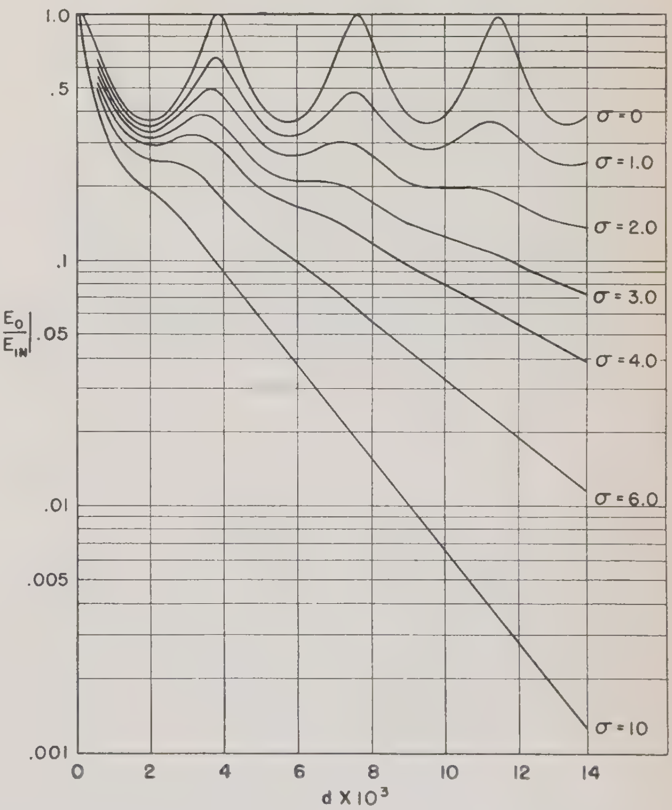

Fig. 8—The magnitude of E_o/E_{in} as a function of conductivity and length. In this case, the frequency is 10^{10} cps, the material is germanium, and dimensions of the waveguide correspond to X band. The cutoff wavelength is assumed to be 4.56×10^{-2} meters.

used independently to estimate conductivity. The reason for choosing the magnitude of E_o/E_{in} (as was done for this paper) lies primarily in its experimental simplicity.

Appendix

Derivation of Waveguide Transmission Equations with Semiconductors Filling a Finite Region

Consider the transmission line shown in Fig. 1. The impedance at the line ab to the left is,

$$Z_{ab} = Z_{02}\left(\frac{Z_{01} + Z_{02} \tanh \Gamma_2 l_2}{Z_{02} + Z_{01} \tanh \Gamma_2 l_2}\right), \tag{4}$$

and

$$E_T = E_{in} r_t, \tag{5}$$

where E_T is the field just adjacent to ab and Z_{02}, E_{in} is the incident field next to ab and Z_{01}, and r_t is the ratio of the two fields. At ab the impedance is Z_{ab}, and at cd the impedance is Z_{01}. At ab, by well-known transmission line theory,

$$E_T = E_o\left(\cosh \Gamma_2 l_2 + \frac{Z_{02}}{Z_{01}} \sinh \Gamma_2 l_2\right). \tag{6}$$

Similarly at cd, by going in the reverse direction,

$$E_o = E_T\left(\cosh \Gamma_2 l_2 - \frac{Z_{02}}{Z_{ab}} \sinh \Gamma_2 l_2\right). \tag{7}$$

Now substituting (5) into (7) we get,

$$\frac{E_o}{E_{in}} = r_t\left(\cosh \Gamma_2 l_2 - \frac{Z_{02}}{Z_{ab}} \sinh \Gamma_2 l_2\right). \tag{8}$$

The term r_t can be obtained by analogous procedures. For a connection between two dissimilar lines,

$$r_t = \frac{2Z_{02}}{Z_{01} + Z_{02}} \cdot \tag{9}$$

However, in the three-dissimilar line case (such as we are discussing), in place of Z_{02} use Z_{ab}. Substituting we find

$$r_t = \frac{2Z_{ab}}{Z_{01} + Z_{ab}} \cdot \tag{10}$$

The next part of the problem is to define each term in these equations in terms of frequency, waveguide dimensions, dielectric constant and conductivity.

The propagation constant in an air filled guide is

$$\Gamma_1 = j\frac{2\pi}{\lambda_1}\sqrt{\left(1 - \frac{\lambda_1}{\lambda_0}\right)^2}, \tag{11}$$

where

$j = \sqrt{-1}$
λ_1 = wavelength in free space,
λ_0 = cutoff wavelength.

In these experiments, λ_1 corresponded to a frequency of 9.549 kMc and $\lambda_0 = 4.56 \times 10^{-2}$ meters (X band). In medium two, the propagation constant is

$$\Gamma_2 = j\frac{2\pi}{\lambda_2}\sqrt{\left[1 - \left(\frac{\lambda_2}{\lambda_0}\right)^2\right] - j\frac{\sigma}{\omega\epsilon}} \cdot \tag{12}$$

where

σ = the conductivity in (ohm-meter),$^{-1}$
$\omega = 2\pi \times 9.549 \times 10^9$ cps,
$\lambda_2 = \lambda_1/4$, which is the wavelength of the radiation in the nonlossy medium,
$\epsilon = 1.41664 \times 10^{-10}$ for germanium[5])
$\lambda_0 = 4.56 \times 10^{-2}$ meters.

Given Γ_2, Z_{02} can be computed by

$$Z_{02} = \frac{j\omega\mu}{\Gamma_2}, \tag{13}$$

where μ is assumed to be 1.257×10^{-6}.

In relation to medium three, which is the same as medium one,

$$Z_{01} = Z_{03} \tag{14}$$

and

$$\Gamma_1 = \Gamma_3. \tag{15}$$

We see from these considerations that for a given thickness, l_2, of the semiconductor slab, the ratio of transmitted field in medium 3 to the incident field in medium 1 will vary with conductivity through (12). Hence for constant l_2 in germanium and at constant frequency,

$$\frac{E_o}{E_{in}} = f(\sigma). \tag{16}$$

In general E_o/E_{in} is complex, containing a magnitude $|E_o/E_{in}|$ and a phase shift term ϕ.

Acknowledgment

The authors are indebted to Drs. L. Hatkin and G. Goubau for their discussions and advice during the course of these experiments, to C. LoCascio for the experimentation, and to Mrs. M. Tate for the application of the computer to much of the theoretical data.

[5] R. D. Middlebrook, "Introduction to Junction Transistor Theory," John Wiley & Sons, Inc., New York, N. Y., p. 292; 1957.

Accuracy and Limitations of the Resistor Network Used for Solving Laplace's and Poisson's Equations*

J. R. HECHTEL†, SENIOR MEMBER, IRE, and J. A. SEEGER†, MEMBER, IRE

Summary—**A resistor network developed for the solution of electron-optical problems is described. New, improved methods for simulating arbitrary boundary conditions with high accuracy are given. The sources of error are discussed. The error is defined by the dislocation of an equipotential, as determined with the network, from its correct position. It is measured in units of the distance between two nodes of the network, called an "interval." For a homogeneous field, the maximum error is below one hundredth of an interval. In most practical problems, the maximum error is well below one tenth of an interval.**

I. Introduction

THERE are many problems in physics and engineering which are described by Laplace's and Poisson's Equations. As analytical methods can be applied only in a few simple cases, these problems are generally solved with the help of physical analogs, such as, for example, the electrolytic tank. Hogan demonstrated that this problem can also be solved by means of a resistor network [1]. In spite of several advantages, the network method as yet has not become very popular. This situation is attributable to the fact that these advantages are not well enough known. An individual not familiar with the theory and the practical use of this instrument usually underestimates its accuracy and resolution. The emphasis of this paper is on the precise simulation of boundary conditions and on the accuracy of the network.

II. Description of the Resistor Network

The network (Fig. 1) developed for the solution of electron-optical problems is of the axially symmetric type. The values of resistors are chosen from the following relations. As seen in Fig. 2, the resistors arranged in the axial (z) direction at a distance

$$r = nh \ (n = 1, 2, 3, \cdots)$$

from the axis have values

$$R_z = \frac{R_0}{n} \cdot \tag{1}$$

The resistors in r direction arranged between $r = nh$ and $r = (n+1)h$ are

$$R_r = \frac{R_0}{n + 1/2} \cdot \tag{2}$$

The resistors on the axis where (1) is invalid, are

$$R_{\text{axis}} = 8R_0. \tag{3}$$

These formulas are derived in the Appendix

The network has 50 units in the z direction and 25 units in the r direction. With respect to resolution and accuracy, a large number of elements is desirable. However, some compromise is necessary because of the factors of size, cost, time for construction, and convenience of use. With $R_0 = 25$ kΩ the values of the resistors vary between 200 kΩ on the axis and 1 kΩ at $r = 25$ h (*cf.* Fig. 2). At the edges of the network (for $z = 0$ and $z = 50$ h), the resistors have twice the value given by (2). This gives a reflecting boundary.

The resistors are of the carbon film type, rated at 0.5 watt with a tolerance of ±0.5 per cent in resistance. All the resistors were checked individually.

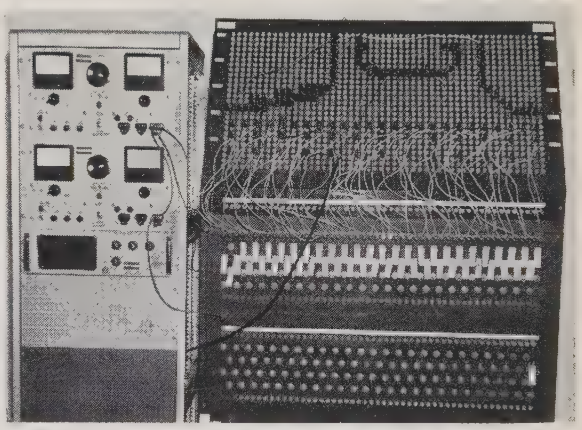

Fig. 1—Resistor network with space-charge simulation.

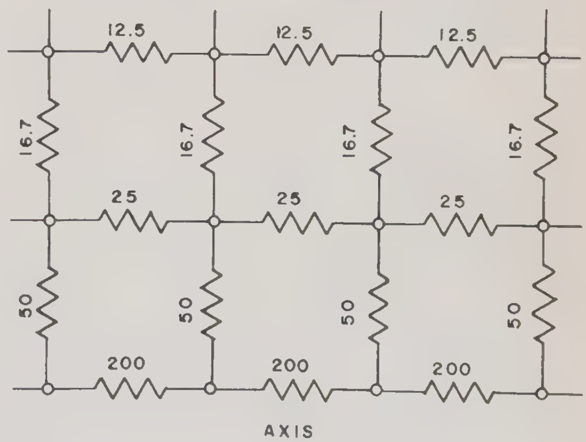

Fig. 2—Arrangement of resistors in a network for axial symmetry.

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III. Operation of the Resistor Network

A. Simulation of Boundary Conditions

Electrodes which do not coincide with the nodes of the network can be simulated by introducing shunt resistors. This has the effect of electrically shifting the position of a node to the desired position. For simplicity in the following calculations, all resistors in the network are assumed to have the same value R. First consider a straight electrode, as shown in Fig. 3. As the field in the vicinity of this electrode is homogeneous, the following relations between the potentials exist:

$$V_1 - V_0 = V_0 - V_3 \tag{4}$$

$$V_e - V_0 = a(V_0 - V_2). \tag{5}$$

To fulfil (5), the resistor representing the interval with the reduced length ah must be changed. The new value may be xR where x is an unknown factor. Applying Kirchhoff's Law, we have

$$\frac{V_1 - V_0}{R} + \frac{V_2 - V_0}{R} + \frac{V_3 - V_0}{R} + \frac{V_e - V_0}{xR} = 0. \tag{6}$$

Combining (4)–(6), we obtain

$$x = a. \tag{7}$$

This means that the resistance has to be reduced in the same ratio as the length of the interval represented by the resistor. To reduce a resistance from R to xR, a shunt resistor

$$R_{sh} = \frac{x}{1 - x} R \tag{8}$$

is added. A second case is possible as shown in Fig. 4, where the distances of a point with potential V_0 from the electrode are ah in z direction and bh in r direction. If the resistors representing the intervals ah and bh are xR and yR, the following equation can be derived in a similar way as before:

$$\frac{1}{x} + \frac{1}{y} = \frac{1}{a} + \frac{1}{b}. \tag{9}$$

Although (9) is satisfied by an infinite number of values of x and y, the most convenient method is to leave one resistor unchanged. With $y = 1$, the condition for x from (9) becomes

$$x = \frac{ab}{a - ab + b}. \tag{10}$$

Fig. 5 shows an example for the arrangement of shunt resistors in the network. Shunts 1, 2, 3 and 5 must be calculated from (7) and (8), and shunt 4 from (10) and (8). Using this method, it is obvious that only one side of an electrode is represented correctly. The backside of the electrode is identical with the broken line given by the electric short between the nodes of the network.

The case of a curved electrode is shown in Fig. 6.

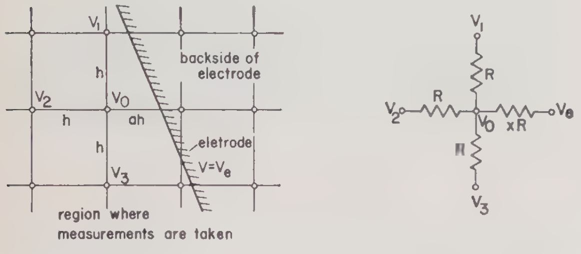


Fig. 3—Simulation of a straight electrode, first case.

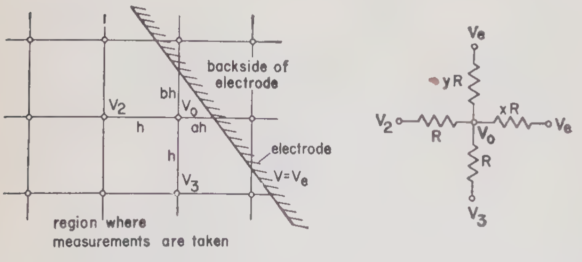


Fig. 4—Simulation of a straight electrode, second case.

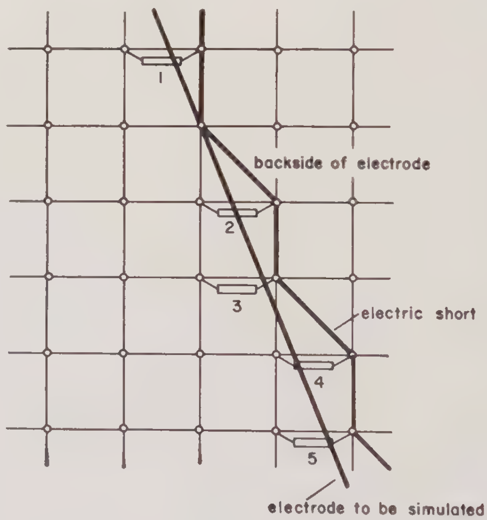


Fig. 5—Arrangement of shunt resistors for simulating a straight electrode

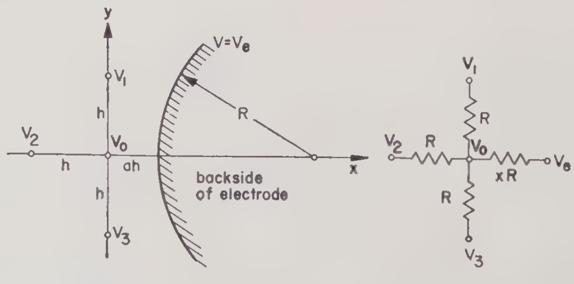


Fig. 6—Simulation of a curved electrode.

Expanding the potential V into a Taylor series and neglecting terms of 3rd and higher order, the following equations can be written

$$V_1 = V_0 + \left(\frac{\partial V}{\partial y}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial y^2}\right)_0 h^2 \tag{11}$$

$$V_2 = V_0 - \left(\frac{\partial V}{\partial x}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial x^2}\right)_0 h^2 \tag{12}$$

$$V_3 = V_0 - \left(\frac{\partial V}{\partial y}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial y^2}\right)_0 h^2 \tag{13}$$

$$V_e = V_0 + \left(\frac{\partial V}{\partial x}\right)_0 ah + \frac{1}{2}\left(\frac{\partial^2 V}{\partial x^2}\right)_0 (ah)^2. \tag{14}$$

As the electrode which is to be simulated represents an equipotential, its radius of curvature is related to $\partial V/\partial x$ and $\partial^2 V/\partial x^2$ by the equation[1]

$$R = \frac{\partial V/\partial x}{\partial^2 V/\partial x^2}\cdot \tag{15}$$

The resistor representing the reduced interval ah is changed to xR, as in the case of a straight electrode.

Applying Kirchhoff's Law, we have

$$\frac{V_1 - V_0}{R} + \frac{V_2 - V_0}{R} + \frac{V_3 - V_0}{R} + \frac{V_e - V_0}{xR} = 0. \tag{16}$$

By combining (11)–(16) with Laplace's Equation

$$\frac{\partial^2 V}{\partial x^2} + \frac{\partial^2 V}{\partial y^2} = 0,$$

the following expression for x can be found:

$$x = a\,\frac{1 + \dfrac{ah}{2R}}{1 + \dfrac{h}{2R}}\cdot \tag{17}$$

In (17), R is positive if the shape of the electrode is convex as seen from the side where measurements are taken. For a concave electrode, R would be negative. On the axis of an axially symmetric system where the radius of curvature of an equipotential is given by

$$R = \frac{2\partial V/\partial z}{\partial^2 V/\partial z^2}\cdot$$

Eq. (17) takes the form

$$x_{\text{axis}} = a\,\frac{1 + \dfrac{ah}{R}}{1 + \dfrac{h}{R}}\cdot \tag{17a}$$

[1] The derivation of this formula can be found in textbooks on electron optics. See, for example, [12], p. 341.

Eqs. (17) and (17a) can be applied with little loss in accuracy to the case where the center of the osculating circle has arbitrary coordinates.

In Fig. 7, the factor x is plotted vs the radius of curvature R for different values of a. Even for small radii of curvature R, the difference between a and x is small. We remember that for a straight electrode, x was exactly equal to a. This means that in most cases, unless a high degree of accuracy is wanted, the simple rules for simulating a straight electrode can also be applied to curved electrodes.

A set of shunt resistors of certain standard values was formed. These resistors were mounted on Johnson plugs so that they could quickly be added to the network.

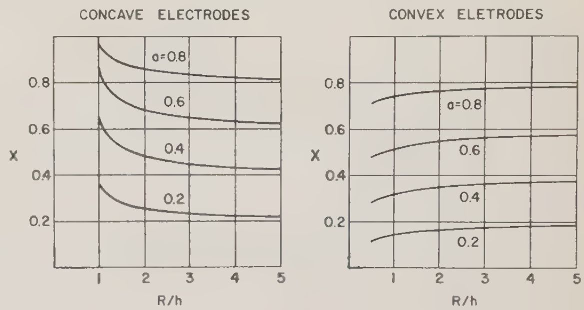


Fig. 7—Auxiliary diagram for simulating curved electrodes.

The values of resistors are that of the RETMA series,

12 15 18 22 27 33 39 47 56 68 82 100,

with a lowest value of 120 Ω and a highest one of 1800 kΩ. With approximately 100 shunt resistors, the electrodes of an average problem can be simulated. In order to further simplify the method, tables were calculated from (8) and (10). The whole work of simulating an electrode is reduced to using the tables, picking out the right shunt resistor from the existing collection, and plugging it into the network. Of course some error is introduced by using the tables and set of fixed resistors instead of the exact values. However, this error in the effective position of an electrode is in all cases below $\frac{1}{10}$ of the interval h.

B. Simulation of Space Charge

As shown in the Appendix, space charge can be simulated by feeding currents into the individual nodes of the network. The space charge density ρ in an electron beam is determined by current density j and the potential V, measured with respect to the cathode. To simulate a current density j, currents

$$I_f = 1.9 \times 10^5 \times \frac{jrh}{R_0\sqrt{V}} \tag{18}$$

are fed into the network within the area occupied by the electron beam. In the above equation, r is the distance of the feed-in points from the axis, h is the distance between two nodes of the network, V is the uncorrected potential at the feed-in point, and R_0 is the characteristic resistance of the network defined by (1)–(3). On the axis, where (18) is invalid, the simulation current is given by

$$(I_f)_{ax} = 0.2375 \times 10^5 \times \frac{jh^2}{R_0\sqrt{V}}\cdot \tag{19}$$

To produce the simulation currents, a source of constant current is desirable. This can be realized closely by a relatively high voltage and series resistors which are large compared to the internal impedance of the network. With the network described in this paper, potentiometers of 5 MΩ and 0.5 MΩ in series with a fixed resistor of 100 kΩ were used for each current.

One hundred such current sources were mounted on the lower board shown in Fig. 1. The simulation currents are determined by measuring the voltages across the 100 kΩ resistors.

To calculate the simulation currents, j and V must first be approximated. This can be done by taking values corresponding to a solution of the problem without space charge. Then by an iterative procedure, a higher degree of accuracy may be reached. In all problems which have been treated, this method has proved rapidly convergent.

As an example of space charge simulation, the potential distribution between parallel grids in the presence of an infinitely extended electron stream with motion normal to the grids shall be treated. An analytical solution of the problem is given in [12]. On the network, the distance between the two grids was chosen to be six intervals. Current density was taken to be 8.2×10^{-3} ma/intervals² (this value is one half the critical current density for which a virtual cathode exists). The voltage of the grids was chosen as $V_g=10$ volts. Simulation currents were calculated from (18) and fed into the network at nodes between the simulated grids. Fig. 8 shows the result of the measurements. With the second approximation, the difference between the measured and analytical potential values is too small to be plotted. The maximum difference is 0.04 V=0.4 per cent of V_g.

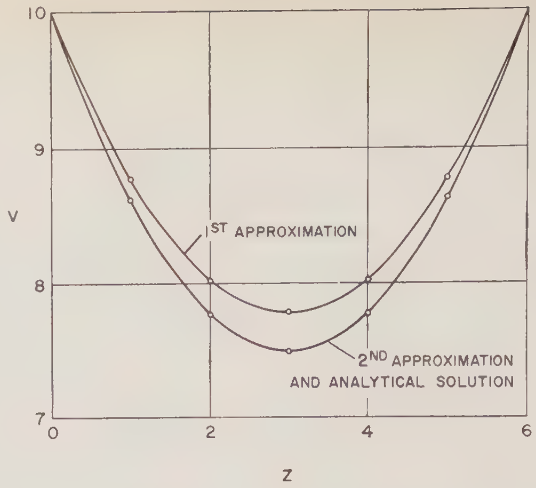


Fig. 8—Potential distribution between parallel grids in the presence of space charge.

C. Measurement Techniques

As the resistor network is operated at relatively low dc voltages, measuring the potential distribution is straightforward. There are two ways used:

1) Potentials are determined by means of a bridge circuit and a galvanometer. With a precision potentiometer or decade resistors, this method is at least as accurate as the network itself. Equipotentials can be found by numerical interpolation, or they can be found directly using an additional potentiometer of high impedance between nodes on each side of the equipotential. This potentiometer performs a linear interpolation of the potential.

2) Potentials are measured directly with a voltmeter of sufficiently high impedance. The authors used for this purpose a four-place digital reading voltmeter, whose accuracy is comparable to the accuracy of the network. This method is especially suitable for numerical methods of ray tracing. The combination of resistor network and a digital computer IBM 709 for ray tracing will be described in a separate paper.

IV. Sources of Error

Two main sources of error exist in a resistor network:

1) Statistical errors, caused by deviations of the individual resistors from their correct values.

2) Systematic errors, caused by the fact that the analogy between the equations for the network and Laplace's Equation is not a perfect one.

Regarding the statistical errors, Liebmann [5] pointed out that the errors of the individual resistors by their random distribution are partially cancelled out. In this way, the accuracy of a resistor network is much higher than might be expected from the tolerances of the resistors.

The value of the resistors also depends somewhat on the power dissipated. The carbon-film type resistors used by the authors decreased in resistance 0.1 per cent for a power of 40 mw. The difference in potential allowed across each resistor is therefore restricted to approximately 30 volts near the axis and 6 volts at $r=25$ h.

To check the long-term stability, several resistors were run at a power level of 0.5 watt. This is more than

10 times the power used in the network for measurements. After 100 hours the change in resistance was less than 0.1 per cent.

Now consider the second source of errors, which is caused by the neglect of terms of 3rd and higher order when the potential is written as a power series of r and z. The error introduced in this way depends on the geometry of the electrodes. It is greatest in the vicinity of sharp edges or electrodes with a small radius of curvature. Numerical values of the error which is to be expected with certain basic electrode forms will be given in the following paragraph.

V. Over-All Accuracy for Some Basic Electrode Shapes

There are two methods of defining the accuracy of a field analog. Consider the simplest case, where the electrodes are two parallel planes. At a certain point between the electrodes, the difference between the measured potential and the correct value is δV. The simplest way of defining the error at this point is to express the deviation δV as a fraction of the voltage V between the two electrodes. However, this definition of the error has a serious disadvantage as will be shown.

Let us assume that the original distance of the two parallel planes is d_1, and the voltage between them V_1. Now the distance between the plates will be increased from d_1 to d_2 by shifting one of the plates parallel to its original position. If the voltage between the plates is increased in the same ratio as the distance, *i.e.*, if $V_2/V_1 = d_2/d_1$, the field between the plates remains constant and the difference δV between the measured and the correct value of potential will be unchanged. However, the error as defined by $\delta V/V$ has changed and finally depends on the distance of electrodes. The second method of defining accuracy is independent of the distance between electrodes. A certain deviation δV in potential from the correct value can also be interpreted as a dislocation of an equipotential, compared to its correct position. If the dislocation, measured in the direction of the electric field, is δS, the relation

$$\delta S = \frac{\delta V}{dV/dS} = \frac{\delta V}{\text{grad } V} \tag{20}$$

is valid. In the example with the parallel planes, the quantity δS for a certain point of the network is independent of the applied voltage and of the distance between electrodes. As the new definition of error given by (20) offers the same advantages in connection with other more complicated electrode shapes, it will be used throughout the rest of this paper. The dislocation of an equipotential is thereby measured in units of the distance between two adjacent nodes of the network, called "intervals." The following cases have been tested and compared to the rigorous theoretical solution.

A. Parallel Planes

The average error is 8×10^{-4} intervals, the maximum error 5×10^{-3} intervals. The only error is in this case caused by the statistical deviations of the resistors in the network from their correct values.

B. Coaxial Cylinders

The theoretical potential distribution is given by

$$V = A + B \ln r. \tag{21}$$

The error depends on the radius of the inner cylinder. For small radii ($r_i<2$) it is mainly of systematic nature; for larger radii, it is a statistical one. For $r_i=1$ the maximum error is 0.02 interval; for $r_i \geqq 2$ the maximum error is 0.01 interval.

If the axis of the network is used as an electrode, it is equivalent to a cylinder of a radius of $\frac{1}{8}$ of an interval.

C. Concentric Spheres

Theoretically the potential is given by

$$V = A + \frac{B}{r}\cdot \tag{22}$$

Similar to the coaxial cylinders, the error depends on the radius of the inner sphere. It is systematic and relatively large for a small radius. The maximum error is

0.1 interval with $r_i = 1$

0.03 interval with $r_i = 2$

0.01 interval with $r_i = 5$

A single point on the axis produces a field similar to that of a sphere with a radius of 0.3 interval.

D. Surface of Revolution Generated by Hyperbolas

With a potential distribution of the form

$$V = z^2 - \frac{r^2}{2}, \tag{23}$$

the equipotentials are hyperbolas. As both $\partial^3 V/\partial z^3$ and $\partial^3 V/\partial r^3$ are equal to zero, no systematic error should be present in the results obtained with the network. The maximum observed error was 0.01 interval.

E. Wedges of 45° and 90°

The potential in the vicinity of a wedge, formed by two straight lines intersecting at an angle α is given in polar coordinates ρ, ϕ by

$$V = A\rho^{\pi/\alpha} \sin \frac{\pi\phi}{\alpha} + B. \tag{24}$$

To simulate this field on a network of the described type, the wedge must be far enough away from the axis, so that the variation of the resistors in the network with r is negligible. With a 45° wedge ($\alpha = 7/4\ \pi$) the maximum error is 0.2 interval. With a 90° wedge ($\alpha=3/2\ \pi$) the maximum error is 0.1 interval.

F. Semi-Infinite Plane

Making $\alpha = 2\pi$, (24) gives the potential in the vicinity of the edge of a semi-infinite plane. The maximum error observed with the network is 0.2 interval. The errors under this and the above section are mainly of systematic nature.

G. 90° Corner

With $\alpha = \pi/2$, (24) describes the potential in the interior of a 90° corner. This type of field can be simulated with great accuracy on the network. The maximum error is only 0.003 interval.

The errors observed in the different cases are summarized in Fig. 9, which also shows the arrangement of electrodes on the network. The largest error of 0.2 interval occurs in the vicinity of the semi-infinite plane and of the 45° wedge. Both cases are of little practical importance. With average problems, the error will always be below one tenth of an interval. It is interesting to compare the accuracy of the resistor network to that of an electrolytic tank. According to our knowledge, only two authors claim an accuracy of better than 1 per cent for the tank. Theile and Himpan [10] measured the potential distribution between parallel plates. Under optimum conditions they found an error in potential of 0.5 per cent of the applied voltage. In simulating this case on the network, the maximum error in potential is only 0.015 per cent, if the whole length of the network is used.

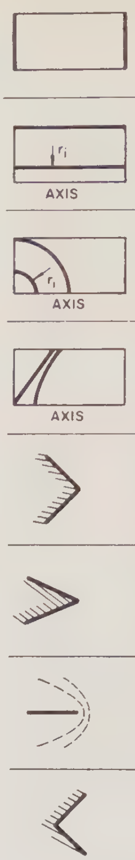
Einstein [11] studied the field between two coaxial cylinders. In the presence of a meniscus between the electrolyte and the electrodes, the error in potential was 1.2 per cent. If the meniscus could be avoided by making the surface of the electrolyte flush with the edge of the electrodes, the error was reduced to 0.2 per cent. With the same ratio of radii (6:1), the maximum error in potential on the network was 0.07 per cent.

An objection which has been raised against the resistor network is that its long-term stability is doubtful. To check this point, measurements were repeated after a period of one year. No change in accuracy was observed.

Conclusions

As an analog computer for solving Laplace's and Poisson's Equations, the resistor network offers several advantages. In comparable cases, its accuracy is higher than that of the best electrolytic tanks. Boundary conditions can be easily set up or changed. Measuring techniques are very simple. The resistance of the network is exactly known and constant in time, an important point in solving Poisson's Equation. A resistor network can be built in relatively short time with a minimum of design and machine shop work.

The spatial resolution of a resistor network of moderate size (25×50 units) is limited. However, experience has shown that the resolution is high enough for all practical purposes.



ELECTRODE CONFIGURATION	MAXIMUM ERROR (INTERVALS)
PARALLEL PLANES	0.008
COAXIAL CYLINDERS	0.02 WITH $r_i = 1$ 0.005 $r_i = 2$
CONCENTRIC SPHERES	0.1 WITH $r_i = 1$ 0.03 $r_i = 2$ 0.01 $r_i = 5$
HYPERBOLAS	0.01
90°-WEDGE	0.1
45°-WEDGE	0.2
SEMI-INFINITE PLANE	0.2
INTERIOR OF 90°-CORNER	0.003

Fig. 9—Accuracy of the resistor network for simple electrode configurations.

Appendix

For an axially symmetric field, Poisson's Equation in cylindrical coordinates r, z is of the form

$$\frac{\partial^2 V}{\partial r^2} + \frac{1}{r}\frac{\partial V}{\partial r} + \frac{\partial^2 V}{\partial z^2} = -\frac{\rho}{\epsilon_0}. \tag{25}$$

Now consider five points arranged in the form of a star as shown in Fig. 10. To find a relation between the potentials at these points, V is developed into a power series.

$$V_1 = V_0 - \left(\frac{\partial V}{\partial r}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial r^2}\right)_0 h^2 - \frac{1}{6}\left(\frac{\partial^3 V}{\partial r^3}\right)_0 h^3 + \cdots \tag{26}$$

$$V_2 = V_0 + \left(\frac{\partial V}{\partial z}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial z^2}\right)_0 h^2 + \frac{1}{6}\left(\frac{\partial^3 V}{\partial z^3}\right)_0 h^3 + \cdots \tag{27}$$

$$V_3 = V_0 + \left(\frac{\partial V}{\partial r}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial r^2}\right)_0 h^2 + \frac{1}{6}\left(\frac{\partial^3 V}{\partial r^3}\right)_0 h^3 + \cdots \tag{28}$$

$$V_4 = V_0 - \left(\frac{\partial V}{\partial z}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial z^2}\right)_0 h^2 - \frac{1}{6}\left(\frac{\partial^3 V}{\partial z^3}\right)_0 h^3 + \cdots \tag{29}$$

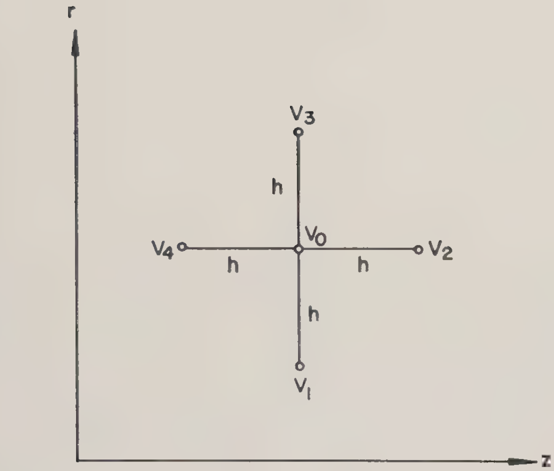


Fig. 10—Arrangement of points used in analysis of network.

Neglecting terms of 3rd and higher order and combining (25)–(28), we obtain

$$V_0 = \tfrac{1}{4}(V_1 + V_2 + V_3 + V_4) + (V_3 - V_1)\frac{h}{8r} + \frac{\rho}{4\epsilon_0}h^2. \tag{30}$$

Fig. 11 shows a section of the network with five nodes arranged in the same way as before. The resistors connecting the center point to its neighboring points are tentatively chosen

$$R_1 = \frac{R_0}{r/h - 1/2} \tag{31}$$

$$R_2 = R_4 = \frac{R_0}{r/h} \tag{32}$$

$$R_3 = \frac{R_0}{r/h + 1/2}, \tag{33}$$

where $r/h = 1, 2, 3, \cdots$ is an integer.

A relation between the potentials can be found by applying Kirchhoff's Law. With a current I_f fed into the center point, we have

$$I_f = \frac{V_0 - V_1}{R_1} + \frac{V_0 - V_2}{R_2} + \frac{V_0 - V_3}{R_3} + \frac{V_0 - V_4}{R_4}. \tag{34}$$

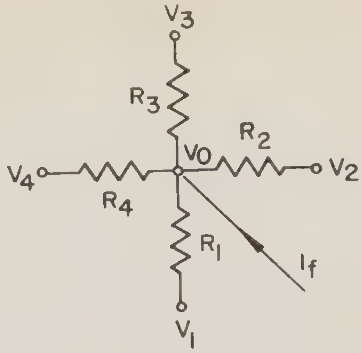


Fig. 11—Section of the network corresponding to Fig. 10.

Combining (31)–(34), we obtain

$$V_0 = \tfrac{1}{4}(V_1 + V_2 + V_3 + V_4) + (V_3 - V_1)\frac{h}{8r} + \frac{I_f R_0 h}{4r}. \tag{34a}$$

A comparison between (30) and (34a) shows that both are identical if

$$\frac{\rho h^2}{4\epsilon_0} = \frac{I_f R_0 h}{4r}$$

or

$$I_f = \frac{\rho h r}{\epsilon_0 R_0}. \tag{35}$$

This means that the potential distribution in a resistor network satisfies Poisson's Equation if the resistors are chosen according to (31)–(33). The space-charge density ρ is thereby simulated by a current I_f, which is fed into each node of the network.

The current density j and the velocity v of electrons can be used to obtain ρ.

$$\rho = \frac{j}{v}. \tag{36}$$

Substituting $v = \sqrt{2e^{v}/m}$, (35) finally can be written

$$I_f = \frac{jhr}{\sqrt{\frac{2e}{m}V}\,\epsilon_0 R_0}$$

$$I_f = 1.9 \times 10^5 \times \frac{jrh}{R_0\sqrt{V}}. \tag{37}$$

If the center point lies on the axis where $V_3 = V_1$ and $r = 0$, the expression $(V_3 - V_1)/r$ in (30) is indeterminant. The equation for the simulation currents is derived in a different manner. Referring to Fig. 12, the potentials V_2, V_3 and V_4 can be written as

$$V_2 = V_0 + \left(\frac{\partial V}{\partial z}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial z^2}\right)_0 h^2 + \cdots \tag{38}$$

$$V_3 = V_0 + \left(\frac{\partial V}{\partial r}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial r^2}\right)_0 h^2 + \cdots \tag{39}$$

$$V_4 = V_0 - \left(\frac{\partial V}{\partial z}\right)_0 h + \frac{1}{2}\left(\frac{\partial^2 V}{\partial z^2}\right)_0 h^2 + \cdots \tag{40}$$

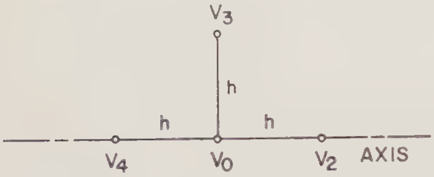


Fig. 12—Arrangement of points used in analysis of network on the axis.

From symmetry,

$$\left(\frac{\partial V}{\partial r}\right)_0 = 0 \cdot \tag{41}$$

Near the axis $\partial V/\partial r$ is given by

$$\frac{\partial R}{\partial r} = \left(\frac{\partial V}{\partial r}\right)_0 + \left(\frac{\partial^2 V}{\partial r^2}\right)_0 r + \cdots \tag{42}$$

By combining (38)–(42) with Poisson's Equation, we find

$$V_0 = \tfrac{1}{6}(V_2 + 4V_3 + V_4) + \frac{\rho h^2}{6\epsilon_0} \cdot \tag{43}$$

Now consider the potential distribution with the same arrangement of points in the network (Fig. 13). By choosing the resistors on the axis to have values of $8R_0$ and applying Kirchhoff's Law, we obtain the following expression for V_0,

$$V_0 = \tfrac{1}{6}(V_2 + 4V_3 + V_4) + \tfrac{4}{3}(I_f)_{ax}R_0. \tag{44}$$

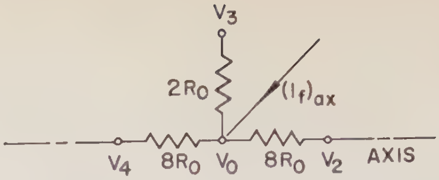


Fig. 13—Section of the network corresponding to Fig. 12.

A comparison of (43) and (44) shows that they are identical if

$$(I_f)_{ax} = \frac{\rho h^2}{8\epsilon_0 R_0} \tag{45}$$

$$(I_f)_{ax} = 0.2375 \times 10^5 \times \frac{jh^2}{R_0\sqrt{V}} \tag{45a}$$

Eq. (45a) gives the current to be fed into a node on the axis in simulating an electron stream with a current density j.

Bibliography

[1] T. K. Hogan, "A general experimental solution of Poisson's Equation for two independent variables," *J. Inst. Engrs.* (Australia), vol. 15, pp. 89–92; April, 1943.
[2] D. C. de Packh, "A resistor network for the approximate solution of the Laplace Equation," *Rev. Sci. Instr.*, vol. 18, pp. 798–799; October, 1947.
[3] S. C. Redshaw, "An electrical potential analyzer," *Proc. Inst. Mech. Engrs.*, vol. 159, pp. 55–62; 1948.
[4] G. Liebmann, "Precise solution of partial differential equations by resistance networks," *Nature*, vol. 164, pp. 149–150; July, 1949.
[5] G. Liebmann, "Solution of partial differential equations with a resistance network analogue," *Brit. J. Appl. Phys.*, vol. 1, pp. 92–103; April, 1950.
[6] G. Liebmann, "Electrical analogues," *Brit. J. Appl. Phys.*, vol. 4, pp. 193–200; July, 1953.
[7] R. Hechtel, "Ein Widerstandsnetzwerk zur Loesung der Poissonschen Gleichung," *Die Telefunken-Roehre*, pp. 38–46; February, 1955.
[8] G. Čremosňik and M. J. O. Strutt, "Bestimmung des Raumladungsfeldes, der Raumladungskapazitaeten und der Kennlinie einer ebenen Triode mit Hilfe eines Widerstandsnetzmodells mit Stromquellen," *Arch. elek. Uebertragung*, vol. 11, pp. 63–75; February, 1957.
[9] G. Čremosňik, A. Frei, and M. J. O. Strutt, "New applications of impedance networks as analog computers for electronic space charge and for semiconductor diffusion problems," Proc. IRE, vol. 46, pp. 868–877; May, 1958.
[10] R. Theile and J. Himpan, "Die Auswahl des Elektrodenmaterials für die Feldbildaufnahme im elektrolytischen Trog," *Die Telefunken-Roehre*, vol. 18, pp. 50–57; June, 1940.
[11] P. A. Einstein, "Factors limiting the accuracy of the electrolytic plotting tank," *Brit. J. Appl. Phys.*, vol. 2, pp. 49–55; February, 1951.
[12] K. R. Spangenberg, "Vacuum Tubes," McGraw-Hill Book Co., Inc., New York, N. Y., p. 256; 1948.

Optimum Capacitor Charging Efficiency for Space Systems*

P. M. MOSTOV†, J. L. NEURINGER†, AND D. S. RIGNEY†

Summary—**In space systems the efficient utilization of energy can be critical. Many pulsed-operation devices, *e.g.*, a plasma engine, involve charging a capacitor bank C periodically. The efficiency of energy transfer η to an initially uncharged bank when a dc-source voltage is applied through a resistance R, with inductance L assumed zero, is limited to 50 per cent even if R varies *arbitrarily* during the charging process. If $L>0$, η can be made to approach 100 per cent by charging in a periodic mode and terminating at the end of the first $\frac{1}{2}$-cycle. The requisite L, which is a function of R, C and charging time T, can be formidable and its weight large, often being excessive for space applications. To provide guide lines in the selection of practical voltage shapes, the techniques of the Calculus of Variations are used to derive a series of general theorems for the "perfect" *time-shaped* source voltages that optimize η when the delivered energy, L, C and T are fixed. Four modes of prescribing R, as a function of time t and/or current i, are treated: 1) Constant R; the key condition is constant i for the full charging time allowed; the voltage is a modified "elevated ramp"; $\eta_{opt}=1/(1+2RC/T)$, can approach 100 per cent and is *independent* of L. 2) $R(t)$. 3) $R(i)$. 4) $R(t, i)$. The η_{opt} for each case, showing the highest theoretically possible, are useful as a basis for an efficiency figure of merit. Non-zero initial bank voltages often lead to improved η's, and are compatible for use with a plasma accelerator stage where they also imply improved energy utilization factors. Promising voltage shapes, approximating the ideal, are synthesized.**

I. Introduction

In an ever increasing number of power systems, as well as those combining power and communication functions, it is of paramount importance to optimize the *efficiency of energy transfer* η, the ratio of the energy delivered to the load to that given up by the source. Such an optimization procedure is distinctly different from the usual one in most communication systems, where a maximum absolute power transfer is generally aimed for without much regard for the efficiency. In fact, in the usual maximum power transfer situations, η is only about 50 per cent.

Unfortunately, a loss of energy is generally reflected in the weight of the energy source. For naval, or even ground applications, this weight can often be somewhat of a problem, but for aeronautical, and especially astronautical applications, it can be quite crucial in its effect on the total system. This is especially true for those propulsion systems in which most of the energy is supplied to, rather than being generated in, the propellant [2].

Many pulsed-operation devices, *e.g.*, Republic's plasma engine [2], [3], involve charging a condenser periodically. A simplified schematic of a typical LCR series-feeding network is given in Fig. 1. The energy source S charges the capacitor bank C by means of the current $i(t)$ through the resistance $R(t)$ and inductance L, t denoting the time. In such a stage the bank plays the role of a load. The bank voltage and charge are $v(t)$ and $q(t)$, respectively, and the source voltage is $e(t)$. At the completion of the charging time T, the bank discharges in a relatively short period of time, *i.e.*, in a pulse fashion, into the next stage. Applying Kirchhoff's law,

$$e(t) - i(t)R(t) - L\frac{di}{dt} - v(t) = 0; \tag{1}$$

$$v(t) = v_0 + \int_0^t i(\tau)d\tau/C,$$

v_0 being its initial value. If a particular $e(t)$ is applied to a given LCR with specified initial conditions, all subsequent behavior is determined. Both the initial current and v_0 will be taken as zero, but the latter will be allowed to be different from zero in Section IV.

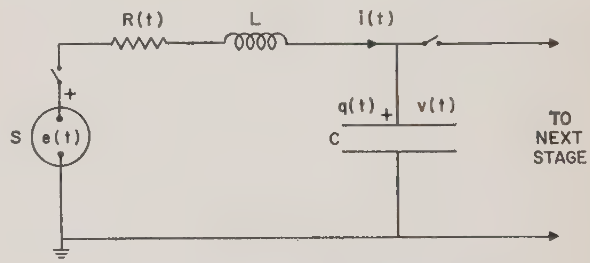


Fig. 1—Schematic of the charging circuit.

Though dc-source voltage schemes can lead to high η's in some regimes (Section II), it may not be practical to use them in many cases, since there are often inherent losses and critical timing requirements in both their creation and application; they also frequently involve heavy inductances. To provide a guide for choosing voltage waveforms that are both efficient and realistic, a series of basic theorems on the voltage shapes that optimize η are derived (Section III) for a variety of conditions resembling practical situations often encountered. For most of the cases treated, η_{opt} is found to be a function of T/RC only, *i.e.*, *independent* of L. Neither the minimum irreducible L, which almost invariably means an aperiodic response to a dc and a 50 per cent η maximum, the large L often required for $\frac{1}{2}$-cycle charging, nor any special L that may be needed for compatibility with other facets of a given design,

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need be a direct limitation on the η attainable. Thus by proper *voltage shaping* it is possible to achieve high η's without stringent restrictions on L and attendant weight problems. η_{opt} for each case, showing the highest theoretically achievable, can serve as the basis for an efficiency figure of merit. The effect of a non-zero initial bank voltage is considered (Section IV); improved η's, some approaching 100 per cent when $v_0 \rightarrow V$ (the final voltage), are shown possible. Fortunately, the $v_0 > 0$ condition is compatible with coupling to a plasma accelerator stage, where it tends to imply efficient energy utilization [7]. Finally, based upon the theorems developed, typical promising, and perhaps more practical, voltage shapes approximating the "perfect" one are synthesized (Section V).

II. DC-Source Voltage

The case where $e(t) = \text{constant}$, a common well-known arrangement, can serve as a useful reference case, as well as help illustrate the basic ideas.

If R and C are constant, and if additionally $L = 0$, η is well known to be

$$\eta_{dc} = \tfrac{1}{2}(1 - e^{-T/RC}), \tag{2}$$

a monotonically increasing function of the single parameter T/RC (number of RC time constants used during the charging time). η vs T/RC for this case is plotted in Fig. 2, in which several other cases to be discussed below are also included for comparison. The maximum value of η attained for the present case is 50 per cent; that such a limitation can make the bank charging stage a serious weak link in many systems has been emphasized by several writers [4, 8].

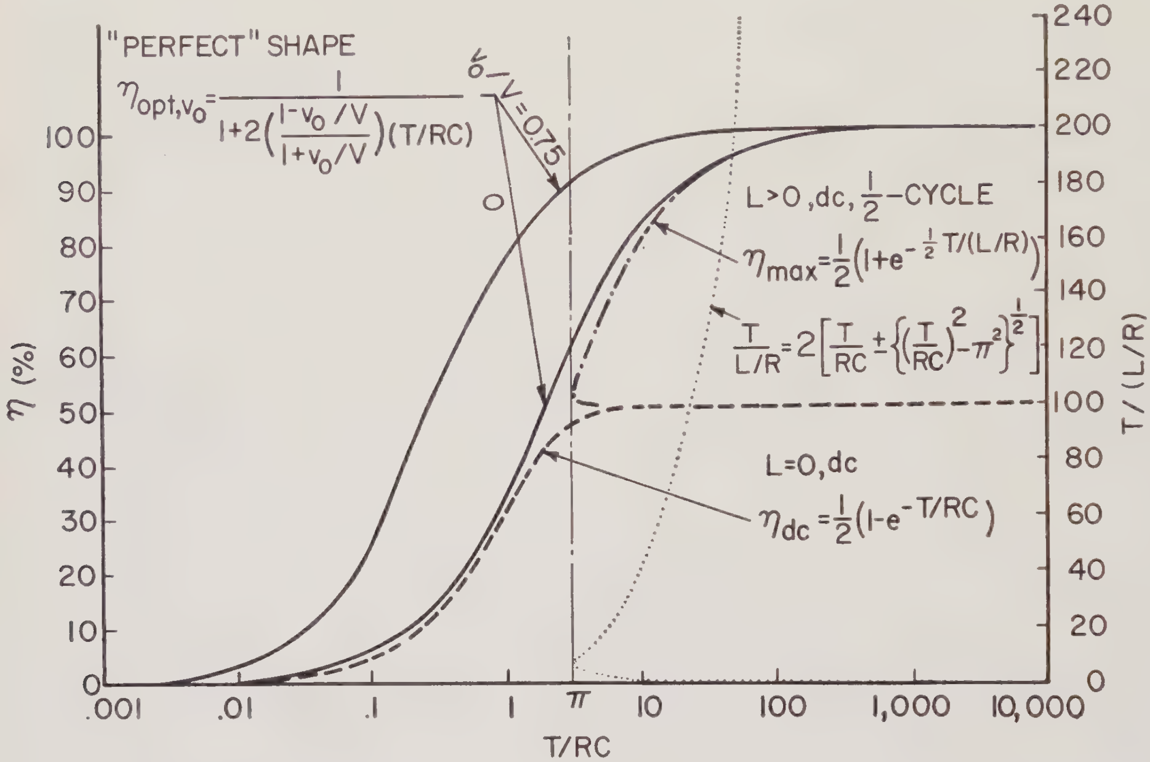

Fig. 2—η vs T/RC for

$L = 0$, dc,
$L > 0$, dc, $\frac{1}{2}$-cycle.
Any L, perfect shape: $v_0/V = 0$, 0.75.

Also $T/(L/R)$ vs T/RC at the $\frac{1}{2}$-cycle time for the $L > 0$, dc, under-damped mode.

This 50 per cent limitation can be shown to govern even if R *varies arbitrarily* during the charging process [1]. Thus, constant-voltage schemes that attempt to circumvent this inherent limitation by varying R are doomed to failure, *e.g.*, those constant-current generators that use a dc-voltage source and vary the effective internal R. However, as shown below, if a constant i is attained by *tailoring* the source voltage properly, this 50 per cent limitation no longer applies; in fact, almost 100 per cent efficiency is possible.

When $L>0$, η can be made substantially larger than 50 per cent under some conditions. As is well known, the current response mode can be aperiodic, critical or periodic. For the first two, it can be shown that $\eta \leq 50$ per cent. For the periodic mode, however, η can exceed 50 per cent under appropriate conditions. The maximum η, attained by arranging to cut off at the end of the first $\frac{1}{2}$-cycle, is

$$\eta_{\max} = \tfrac{1}{2}(1 + e^{-\frac{1}{2}T/(L/R)}), \tag{3}$$

which can also be expressed in terms of only T/RC since

$$\frac{T}{L/R} = 2\left[\frac{T}{RC} \pm \left\{\left(\frac{T}{RC}\right)^2 - \pi^2\right\}^{1/2}\right] \tag{4}$$

must hold ($\frac{1}{2}$-cycle time condition). The L required is *determined* for a given R, C and T; it often can be quite large and its associated weight formidable, especially if its contribution to the total R is to be kept reasonable. There are two branches of $\eta(T/RC)$ [cf. Fig. 2, on which (4) is also plotted to help indicate the required associated L's]. On the higher branch of $\eta(T/RC)$, to which the lower $T/(L/R)$ branch and therefore the *higher* L correspond, η increases from 52.2 to 100 per cent; note for large T/RC, $L \propto T^2/C$. On the lower branch, η decreases from 52.2 to 50 per cent. Eq. (4) can only be satisfied for $T/RC \geq \pi$ if $T/(L/R)$ is to be real. If (4) is not satisfied, (3) does not govern, but substantial η's are still possible provided $T/RC > \frac{1}{4}T/(L/R)$ (under-damped).

Thus, though constant e schemes can imply high η's in some ranges, they are not always usable; *e.g.*, they often involve heavy inductances and critical timing requirements (i is changing most rapidly at the $\frac{1}{2}$-cycle time, so that a small error in timing can lead to undesirable i's, which in turn can mean energy losses, wear and tear on electrodes and switches).

III. Optimization Theorems

We now inquire how to specify e as a function of t so as to improve on the dc-source case with its inherent 50 per cent limitation when $L=0$ and other restrictions when $L>0$, or for that matter on any other e. At first, $L=0$ will be treated with constant R, then $L>0$ will be considered, and eventually R will be allowed to be a prescribed function of t, i, or even both. Thus, to serve as a guide in the selection of waveforms that are both efficient and practical, and to tell us the best possible η we may expect, a series of theorems on the optimization of the efficiency will be derived. Employing the techniques of the Calculus of Variations [5, 6], the "perfect" *time-shaped* e's that optimize η for prescribed LRC, T and delivered energy will be found. It is simplest to recast the original question on e to one on the associated i, and the key condition is most succinctly stated in terms of the i-t characteristic.

A. $L=0$, R Constant

For a given C, specifying the final condenser charge, Q, is equivalent to specifying its energy as well as V, since they equal $Q^2/2C$ and Q/C, respectively. But as

$$Q = \int_0^T i(t)dt, \tag{5}$$

specifying the $i-t$ integral is equivalent to specifying the total energy delivered. Consequently, to optimize η it is sufficient that the resistive losses

$$U_R = \int_0^T i^2 R dt$$

be minimized, subject to (5).

To minimize the area under the i^2R-t curve, subject to a prescribed area under the i-t curve, i should be *constant* over the whole time available to charge. This can be seen by realizing that any slight variation of the flat i-t curve, while keeping the area under it constant, would raise the integrand of the i^2R-t curve, at the point in t where i is raised, *more* than it would lower the integrand at any point where i is lowered an equal amount—any raising of i at one point would necessitate the lowering of i at another point(s) in order to maintain Q constant. Thus, the area under the i^2R-t curve would *increase*. This argument can be generalized to the case where the rise of i at one point is compensated by the lowering of i at more than one point.[1] For constant i, (5) yields $i=Q/T$ (Fig. 3).

That a constant i from $t=0-T$ is indeed the requisite condition can be demonstrated mathematically by

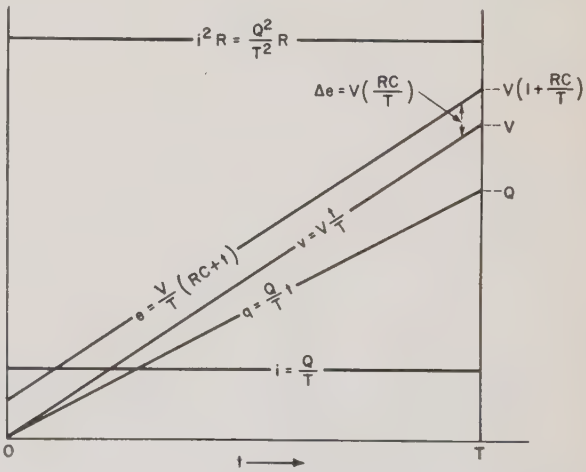

Fig. 3—i, q, v, Δe, e and i^2R vs t for optimum η, when the delivered energy, R, C and T are prescribed constants and $L=0$. When $L>0$, suitable positive and negative pulses must be superimposed on e at $t=0$, T, respectively

[1] It can also be shown that this minimum is the lowest possible one.

using the Calculus of Variations, which is also a potent method to treat more complicated cases (below).

The problem is to find the $i(t)$ which makes U_R a minimum, subject to (5).

$$\delta \int_0^T (i^2R + \lambda i)dt = 0$$

is needed, where λ is a Lagrange multiplier. Since i is the variable whose "path" in t is varied, the applicable Euler-Lagrange equation is

$$\frac{d}{dt}\left(\frac{\partial \mathcal{L}}{\partial i'}\right) - \frac{\partial \mathcal{L}}{\partial i} = 0, \tag{6}$$

where $\mathcal{L}$ denotes the Lagrangian and the prime denotes differentiation with respect to t. As undetermined end points are allowed,

$$\left[\frac{\partial \mathcal{L}}{\partial i'}\delta i\right]_0^T = 0 \tag{7}$$

must hold. Since here $\mathcal{L} = i^2R + \lambda i$ and does not contain i', (7) is satisfied. The d/dt term of (6) vanishes and

$$2iR + \lambda = 0; \tag{8}$$

i.e., $i = -\lambda/2R$, a constant. λ may be determined by (5). As the *lowest* minimum is not always an extremum, finite variations should be used to search for it.

The resistive loss rate is $i^2R = Q^2R/T^2 = \text{constant}$. A constant i feeding C means that q and v rise linearly from 0 to respectively Q and V; *i.e.*, $q = Qt/T$, $v = Vt/T$. Δe, *i.e.*, $(e-v)$, is simply the iR drop and is constant:

$$\Delta e = RQ/T = V(RC/T).$$

The required e is $e = (RC+t)V/T$, *i.e.*, it too is linear in t, always being Δe higher than v. It does not pass through the $e-t$ origin as one might suppose. At $t = T$, $e/V = 1 + RC/T$. The energy given up by the source is

$$U_s = \int_0^T e i dt = \tfrac{1}{2}CV^2(1 + 2CR/T),$$

that received by the bank $\Delta U_c = \frac{1}{2}CV^2$. As $\eta = \Delta U_c/U_s$,

$$\eta_{\text{opt}} = \frac{1}{1 + \dfrac{2}{T/RC}}, \tag{9}$$

a function of the single parameter T/RC (Fig. 2). It increases monotonically with T/RC and approaches 100 per cent!

In the fixed-e scheme for $L=0$, T must be greater than $3RC$ for the bank to be charged to within 95 per cent of the source voltage. Even for $T/RC = 3$, $\eta = 60$ per cent for the perfect voltage scheme; if $T/RC > 3$, η can be made to approach 100 per cent, but would still be limited to a maximum of 50 per cent for the fixed-e case. T/RC may be increased by decreasing R or C, increasing T, or any combination (one or two of these factors may be changed opposite to that just indicated and still achieve an increase in T/RC provided the other factor(s) are changed sufficiently). Naturally they must be integrated to an over-all system design.

B. Constant L

1) R Constant:

The losses that now have to be minimized are not only integrated i^2R losses, but also the magnetic energy at the time that the charging is completed, $\frac{1}{2}Li^2(T)$, since much of such energy is lost in heating the circuit and the switch, radiation, etc., and is consequently unrecoverable. Thus,

$$\int_0^T (i^2R + iLi')dt$$

should be a minimum, subject to (5). Now,

$$\mathcal{L} = i^2R + iLi' + \lambda i,$$

and contains i' explicitly. Since $i(0) = 0$, (7) reduces to $iL\delta i]_T = 0$. Thus either $i(T)$ or $(\delta i)_T$ must be zero; since it is not known *a priori* that $(\delta i)_T$ is zero, $i(T)$ must be.[2] Eq. (6) becomes $d(iL)/dt - \partial(i^2R + iLi' + \lambda i)/\partial i = 0$, which reduces to (8). Consequently i is the constant Q/T for $0 < t < T$, but zero at *both* end points, *i.e.*, no magnetic energy is trapped in the magnetic field at $t = T$. Previously, none was trapped as L was zero.

Fig. 3 turns out to be still valid *except* for Δe and e near $t = 0$, T. During the time i is flat, Δe and e are the same as for the $L = 0$ case. The extrapolation of e does not pass through the e-t origin. Near the end times the e required can be very extreme, since $e = v + iR + Li'$ and i' is infinite. Strictly speaking, the e pulses necessary to bring i up from 0 to Q/T and down to 0 again are "infinite," positive and negative, respectively, and of "infinitesimal" duration. Since it may be impractical to have peaks that are too large, it is of interest to see what the effect of finite square pulses of various heights and widths would be on i and η (cf. Section V).

The energy invested in the magnetic field to build i to $\sim Q/T$ is $\sim \frac{1}{2}L(Q/T)^2$. This energy however is fully recovered when i decreases to 0. Thus, the net magnetic energy change, ΔU_L, is zero. The i^2R loss during the very short rise (ϵ_1) and fall (ϵ_2) times, $\leq R(Q/T)^2(\epsilon_1 + \epsilon_2)$ (the upper bound is estimated on the basis of the full Q/T for the entire ϵ_1 and ϵ_2 intervals), vanishes as $\epsilon_1, \epsilon_2 \to 0$. The *permanent finite* losses are then simply the i^2R loss during $0 < t < T$: Q^2R/T. As

$$\eta = \Delta U_c/(\Delta U_c + \Delta U_L + U_R)$$

$\eta_{\text{opt}} = 1/(1 + 2/(T/RC))$, a function of T/RC *only*, and is identical with (9). Consequently, some of the previous discussion is also applicable here.

In the RC case, the comparison was with the dc

[2] If $i(T)$ were prescribed, $(\delta i)_T$ would be zero and (7) would still be satisfied.

scheme, in which the maximum η was 50 per cent. For $L>0$, only with $T/RC>\frac{1}{4}T(L/R)$ (under-damped) was it possible to have $\eta>50$ per cent when dc is used. The maximum of such η's was found for the certain *finite* charging times; alternatively, if T, R, C, are given, the L needed is *determined. By proper voltage shaping we have shown that it is possible to achieve $\eta>50$ per cent without such severe restrictions on* T and L. The optimum η achievable is independent of L; *i.e., no special* L, with its often very serious weight problem, is necessary when T, R, C are specified. The e needed, however, does depend on L.

Neither the smallest unavoidable L (which, when a dc is applied usually means an over-damped system and therefore a 50 per cent maximum for η), the sizable L frequently needed for efficient $\frac{1}{2}$-cycle charging, nor any other L that may be helpful to an over-all design, need be a bound on the achievable η.

Though η_{opt} exceeds the upper branch of $\eta_{\max}$ by only a small amount in some ranges of T/RC, especially at the higher T/RC's necessary for substantial η's, it should be kept in mind that the requirement of a large L for the $\frac{1}{2}$-cycle condition can become quite severe and untenable for many space designs.

In many cases treating the circuit resistance as approximately constant is very unrealistic, and its variableness must be taken into account. R can vary due to a number of causes, *e.g.*, heating, cooling, effective radiation resistance, effective internal resistances (including those of constant-i generators), skin effects, solid state effects, plasmas, discharges, various nonlinear components and circuits.

2) $R(t)$ as a Prescribed Function of t:

If $R(t)$ is prescribed, R must be interpreted as $R(t)$ up to and including (8), so that $i(t)R(t)=A$, or $i(t)=A/R(t)$, for $0<t<T$ where $A=$constant. Eq. (7) is satisfied by $i(0)$, $i(T)=0$ for finite L's, but if $L=0$, $i(T)$ may be other than zero.

That this current corresponds to a *minimum* U_R can be seen by varying the i-t curve slightly, while maintaining the area under it equal to Q. Suppose at t_1, i is increased to $(i_1+\delta i)$. One way to keep the i-t area constant is to decrease i at another time t_2 by δi to $(i_2-\delta i)$. The net gain to U_R is

$$[(i_1+\delta i)^2R_1 - i_1^2R_1 + (i_2-\delta i)^2R_2 - i_2^2R_2]dt,$$

which reduces to $(\delta i)^2(R_1+R_2)dt$ when the constancy of iR is used; the total change is always positive (unless negative or zero R's are allowed). As t_1, t_2 were chosen at random, the curve is a minimum.

Using (5), $A=QR_H(T)/T$, where

$$1/R_H(t) = (1/t)\int_0^T (1/R(\tau))d\tau$$

is the time-averaged value of $1/R(t)$ over the 0-t interval, *i.e.*, the "harmonic" time average of $R(t)$. As $R(t)$ is prescribed, $R_H(t)$ may be calculated in advance. Thus

$$i(t)R(t) = \frac{QR_H(T)}{T} = \frac{VCR_H(T)}{T} \tag{10}$$

and $i(t)=(Q/T)R_H(T)/R(t)$ $(0<t<T)$. Also

$$v(t) = VR_H(T)t/R_H(t)T.$$

Substituting in (1),

$$e_{0<t<T} = \frac{VR_H(T)C}{T}\left(\frac{1}{R_H(t)C}t + 1 - \frac{L}{R^2(t)}\frac{dR(t)}{dt}\right).$$

For $e(0)$, $e(T)$ the development is similar to that of the constant R case: additional positive and negative pulses are needed at $t=0$, T if $L>0$.

$\Delta U_L=0$: if $L=0$, because $i(0)$, $i(T)$ are finite; if $L>0$, because $i(0)$, $i(T)=0$. Using[3] (10),

$$U_R = Q^2R_H(T)/T.$$

Thus

$$\eta_{\text{opt}} = \frac{1}{1+2/(T/R_H(T)C)},$$

a form similar to (9), but R is replaced by $R_H(T)$. Consequently Figs. 2 and 3, as well as the discussion of (9), are somewhat applicable here and large η's are generally possible. There is one important difference: To increase η, $T/R_H(T)C$ should be increased. Previously, it was possible to accomplish this by varying T, C and/or R; there were *three* independent parameters. Here, however, $R_H(T)$ is a *function* of T. For a given $R(t)$, one may construct $R_H(T)$ and $T/R_H(T)$ in advance. Then one could attempt to maximize $(T/R_H(T))/C$, where $T/R_H(T)$ and C (or alternatively T and C) are considered the variables. On the other hand, one can alter the efficiency possibilities by changing the prescribed $R(t)$ itself. As before, these relationships must be integrated into an over-all system design.

3) $R(i)$ as a Prescribed Function of i:

The resistance a prescribed function of i, $R(i)$, is usually a closer approximation to the actual physical phenomenon than is $R(t)$; *i.e.*, the value assumed by a resistance usually depends more on the i through it than on the particular t that it happens to be, though of course it is actually a function of both. For a wide class of $R(i)$'s the key i condition and η_{opt} are found to be similar to that of the constant R case.

Instead of (8) one obtains

$$2iR(i) + i^2\frac{dR(i)}{di} + \lambda = 0, \tag{11}$$

and (7) is satisfied. Since $R(i)$, and therefore $dR(i)/di$, is prescribed, (11) may, in principle, be solved for i as a function of λ. There can be many values of i for a

[3] The contributions at $t=0$, T to the U_R integral are nil since the integrands at these points are finite.

given λ, but they will be constants, *i.e.*, not functions of t. For each, λ may be determined by (5); *i.e.*, λ is adjusted so that constant i satisfies the equation. As i is constant, it may be taken out of the integral sign and $i=Q/T$. Consequently, all solutions coalesce to a single one.

To find the conditions under which the solution is a minimum, one may evaluate the sum of the $\delta(i^2R)$'s at two random t's. To 2nd order,

$$\delta(i^2R) = \frac{d(i^2R(i))}{di}\delta i + \frac{1}{2!}\frac{d^2(i^2R(i))}{di^2}(\delta i)^2.$$

As $i_1=i_2$, all derivatives are independent of i, and since $\delta i_2=-\delta i_1$, the sum of $\delta(i^2R)$'s is $(\delta i)^2$ times

$$2R + 4i\left(\frac{dR}{di}\right)_{i=Q/T} + i^2\left(\frac{d^2R}{di^2}\right)_{i=Q/T} \tag{12}$$

For such[4] $R(i)$ that (12) is positive, $i=Q/T$ corresponds to a minimum η.

There are some $R(i)$'s that make (12) negative and yet lead to η's that are higher than those for the $i=Q/T$ curve. The $R(i)$ of Fig. 4(A) decreases slowly, except it drops very rapidly when $i \sim Q/T$. If $i=Q/T$ is subjected to *finite* variations while maintaining Q constant [Fig. 4(B)], it is found that curves a, b and c have successively lower i^2R losses. Note that though $R(i)$, a and b are *continuous*, c is not.

For $i=Q/T$, Fig. 3 is valid except for Δe and $e(t)$ near $t=0$, T. Since $R(i)$ is a function of i, the $iR(i)$ drop can be different from the constant R case during the short rise and fall times; in the limit, they effectively vanish when compared to the infinite pulses needed to overcome $-Ldi/dt$. (For non-infinite pulses that cause i to rise to Q/T in a finite time, the $iR(i)$ drop certainly must be considered.) η is given by (9).

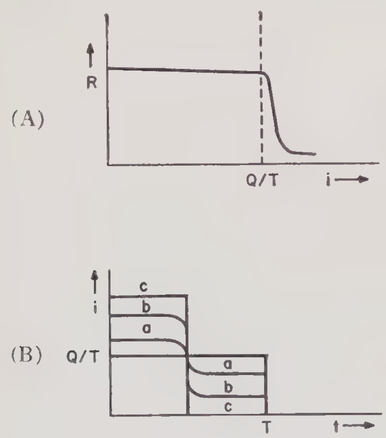

Fig. 4—(A) $R(i)$ decreases slowly, but drops very rapidly when $i \sim Q/T$. (B) Particular finite variations from $i=Q/T$ that have successively lower i^2R losses, while maintaining Q constant.

[4] The requirement that this derivative be positive at all points of the test $i(t)$ is equivalent to the Legendre condition for a minimum ([6], p. 214).

4) *$R(t, i)$ as a Prescribed Function of Both t and i Simultaneously:*

Since $i(t)$ is, *a priori*, an unknown function of t, R is also an unknown function of t even though its dependence on t and i is prescribed. In place of (8),

$$2i(t)R(t, i) + i^2(t)\frac{\partial R(t, i)}{\partial i} + \lambda = 0 \tag{13}$$

governs. For a particular prescribed $R(t, i)$, $\partial R(t, i)/\partial i$ is differentiable in advance so that (13) is an equation for i with λ as a parameter. The solution (there may be many) is a function of t and λ, *i.e.*, $i(t, \lambda)$. For each such, there may be multiple values of λ determined by imposing (5). For each of these cases one may apply the Legendre minimum condition to find which (if any) are minimums; then one may determine which of those is the lowest and whether it is the lowest possible, as the lowest minimum may not be an extremum. For most cases η_{opt} can be readily calculated by the methods used earlier.

IV. Initial Voltage

If the mode of operation is such that the initial bank voltage is greater or less than zero, improved or degraded efficiencies, respectively, compared to when $v_0=0$, will now be shown possible. η can in fact approach 100 per cent as $v_0 \to V$. However, a different-from-zero v_0 also means that not all of the maximum energy that the bank attains is utilized in the discharge process; for a given energy actually used, a bank with a larger energy rating is needed. Naturally, the implications of the greater bank weight that may be required vs the improved η's must be considered in achieving a practical over-all system design.

For $L=0$ and constant R, if the dc voltage is denoted by E, (2) generalizes to

$$\eta_{\text{dc},v_0} = \frac{1}{2}\left\{\left(1-\frac{v_0}{E}\right)(1-e^{-T/RC}) + 2\frac{v_0}{E}\right\},$$

which increases monotonically from v_0/E to $\frac{1}{2}(1+v_0/E)$ as T/RC traverses 0 to ∞; and $\to 100$ per cent as $v_0/E \to 1$. If R varies, it can be shown [1] that η is monotonic in T/RC; and that η for $T/RC=\infty$ is monotonic in v_0/E.

For $L>0$ and constant R, (3) generalizes to

$$\eta_{\max,v_0} = \frac{1}{2}\left[(1+e^{-\frac{1}{2}T/(L/R)}) + \frac{v_0}{E}(1-e^{-\frac{1}{2}T/(L/R)})\right]$$

[(4) is unmodified]. Each branch is monotonic in T/RC. Its minimum and maximum both occur asymptotically as $T/RC \to \infty$ and are $\frac{1}{2}+v_0/E$ (lower branch) and unity (upper branch), respectively. Furthermore, $\eta_{\max,v_0} \to 100$ per cent as $v_0/E \to 1$.

A constant $i=(Q-q_0)/T$, where q_0 is the initial charge, is found to be needed to optimize η when R is a prescribed constant. The required voltage is $e=v_0+(V-v_0)(RC+t)/T$ (an elevated ramp), plus suitable pulses superimposed at $t=0$, T if $L>0$. The optimum η is

$$\eta_{\text{opt},v_0} = \frac{1}{1+2\left(\dfrac{1-v_0/V}{1+v_0/V}\right)\Big/ (T/RC)}, \tag{14}$$

which increases monotonically from 0 to 100 per cent as v_0/V traverses -1 to 1 and is plotted in Fig. 2 for $v_0/V=0$ and 0.75. For $T/RC=3$, it jumps from 60 per cent when $v_0/V=0$ to 90 per cent when $v_0/V=0.75$, a factor of one-half which can be very helpful in many tight, weight-conscious space designs. Should it be desirable, or necessary, to operate at lower T/RC's, the fractional increase is even more dramatic: at $T/RC=0.1$, it exceeds 5. For $v_0=0$, operating at $T/RC<3$ implies that even η_{opt} will be limited to values below 60 per cent, which, in many cases, is impractical except for relatively short missions when the power supply weight tends to be less important than that of the bank. For the longer missions, however, when the power supply weight often tends to be more important than that of the bank, one can still operate with high η at $T/RC<3$ if a substantial v_0 is used.

$\eta_{\text{opt},v_0} \rightarrow 100$ per cent as $T/RC\rightarrow\infty$. For $T/RC>50$ the increment due to $v_0>0$ is not too significant since η already exceeds 90 per cent when $v_0=0$, but for some very long missions it can still be helpful.

When the energy actually used (as opposed to the total energy the bank is called upon to hold) is fixed and V and the energy densities in the bank are restricted to stay within given bounds, the C needed is a function [1] of v_0. Under such circumstances, v_0 and C (and therefore also T/RC) in (14) would not be independently specifiable and the use of the equation would have to be correlated with the C, v_0 relationship. Similarly, a variety of other useful relations may be imposed. It should be noted that (14) must be interpreted with caution. If constraints different from those used in its derivation are employed, a different formula for η_{opt,v_0} can result. The methods illustrated here can handle most of the situations encountered.

V. Promising Practical Shapes

The theorems for the "perfect" source voltage shapes can serve as useful guides in the synthesis of promising, practical voltages. Several typical ones (Fig. 5) are given below. Each resembles the ideal in one or more important features that probably have reasonable chances of being achievable, while minimizing those aspects for which too much of a penalty may be incurred. The η and i characteristics of such basic voltage shapes may be determined analytically, as well as by computers, to help assess the severity of the deviations from the optimum caused by each type of modification. A useful "efficiency merit factor" is the ratio of η to the η_{opt} possible for a given T/RC. For the perfect shapes, the inductive energy U_L is zero at $t=T$: because $i(T)$ is finite when $L=0$; because $i(T)=0$, when $L>0$. Thus for these perfect shapes one need not try to utilize the U_L left as there is none. For practical voltage shapes, however, i is generally not zero at $t=T$, and it may be useful to estimate $U_L(T)$ and consider methods, *e.g.*, crowbarring, to use it where worthwhile. In any event η and efficiency merit factor estimates that ignore such energies serve as useful conservative bounds.

a) *Ramp*, *i.e.*, a slope through the *e-t* origin.
b) *Elevated Ramp*, *i.e.*, a slope *not* through the origin. Though it resembles the perfect shape for the constant R, $L=0$ case, it is not necessarily identical to it since the initial value and slope may need to be adjusted to compensate for the fact that the requisite pulses are not used. If the delivered energy is prescribed, the values of the parameters specifying e may be arranged so as to maximize η.
c) *Elevated Ramp with Finite Pulses*, *i.e.*, an elevated ramp with suitable pulses of *finite* width and height superimposed, rather than the infinitely high and infinitesimally narrow pulses of the perfect shape in the constant R, $L>0$ case.
d) *"Plateaued" Elevated Ramp*, *i.e.*, an e that starts out as an elevated ramp, but eventually becomes flat. Such a modified elevated ramp has the advantage of tending to minimize the cutoff i without resorting to pulses. Furthermore, it can often do this without being critically sensitive to the exact

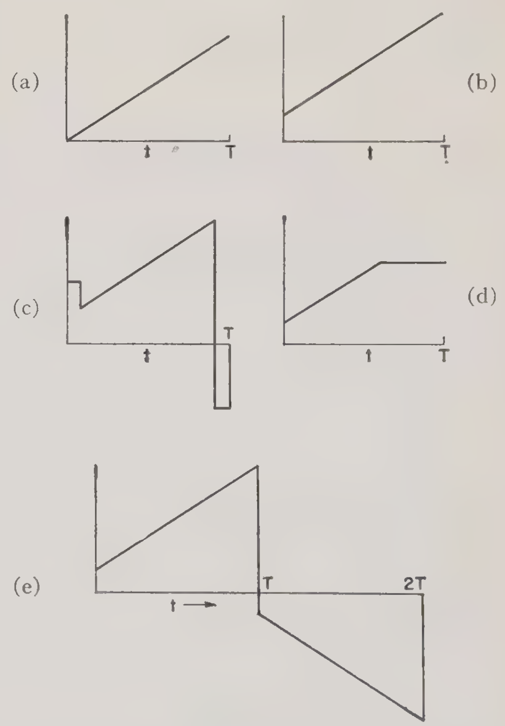

Fig. 5—Several promising voltage shapes: (a) slope, (b) elevated ramp, (c) elevated ramp with finite pulses, (d) "plateaued" elevated ramp, (e) alternating elevated ramp.

time of cutoff. If the plateau portion is continued long enough, i would decrease to zero. This can be seen by realizing that at the time the plateau portion starts, one has the simple problem of a dc e, but with non-zero initial conditions. It is likely that such a scheme would imply extra charging time and tend to erode η. Nevertheless, the minimizing of the cutoff i in a less critically time-sensitive fashion and its concomitant dividends in easing the magnetic energy losses and the switch wear, tear, and reliability problems, if one is used, could be important in some ranges of operation. Should no switch be used, one must of course check that the current drain on the source during the discharge, essentially a short-circuited load, is tolerable.

e) *Alternating Elevated Ramp, i.e.*, for the first of each of two successive T periods the voltage is one polarity and for the second, it is the opposite. (The short discharge time between these two charging periods is not shown.) Note that for the second period, even though the polarity of v_0 is reversed, so is that of V, and the v_0/V ratio remains positive. Consequently, the same increased η (compared to the $v_0/V=0$ case) is possible for both periods. Thus it is not necessary to restrict e to single polarity types, nor to require that the final and initial voltages of the pulse discharge into the load (*e.g.*, a plasma accelerator) have the same polarity as each other; *i.e.*, the voltage polarity at the end of a particular discharge does not have to be the same as that with which the discharge began. All that is needed is that the polarities of the initial and final voltages of a particular *charging* period be the same.

VI. Conclusion

A series of general theorems for the "perfect" time-shaped source voltages that optimize η for a variety of conditions have been derived, and it has been shown that their use minimizes the dependence on heavy inductances. The favorable effect of positive initial bank voltages on η has also been demonstrated. Using the theorems as a guide, promising voltage shapes, approximating the ideal, have been synthesized. The probably high η's attainable with these and other promising shapes are being investigated both theoretically and experimentally. If these shapes can be generated efficiently enough by modified conventional power supplies, a useful efficient generator-feeding network complex would appear to be easily achievable. Otherwise, one of the benefits of the present study is to indicate a gap that novel, unconventional generators should attempt to bridge.

Finally, it should be emphasized that in most of the pulsed plasma accelerators under consideration today, the initial and final voltages of the acceleration stage are identical to the final and initial voltages, respectively, of the charging state—the bank is a common one. Since the energy utilization factor in the acceleration stage is an increasing function [7] of $(v_0/V)^2$, it is possible to achieve increased η's and utilization factors simultaneously. In fact, it is also possible to use waveforms that change polarity in alternate charging times (*e.g.*, shape e), and still achieve an increased η and utilization factor. Thus it turns out that the two stages are compatible for a combined design. An overall optimization of the combined generator, feeding network and plasma accelerator would appear to be a fruitful next step.

In future papers, R as a prescribed function of its previous i-t history, as well as several allied modes of specifying R, L, C, will be treated; *e.g.*, the commonly encountered phenomenon of L varying with i, $L(i)$, and the important case of *mechanically* completing the charging of a capacitor, *i.e.*, $C(t)$. The effect of non-zero initial and final currents will also be given.

Acknowledgment

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Bibliography

[1] P. Mostov, J. Neuringer, and D. Rigney, "Optimum Capacitor Charging Efficiency for Space Systems," Plasma Propulsion Lab., Republic Aviation Corp., Farmingdale, N. Y., PPL-TR-60-2; January 12, 1960. Also cf. PPL 104, 106, 113.

[2] A. Kunen, "The Electromagnetic Pinch Effect Applied to a Space Propulsion System," Plasma Propulsion Lab., Republic Aviation Corp., Farmingdale, N. Y., August 4, 1958.

[3] A. Kunen and W. McIlroy, "The Electromagnetic Pinch Effect for Space Propulsion," Plasma Propulsion Lab., Republic Aviation Corp., Farmingdale, N. Y., PPL No. 116; August, 1958.

[4] B. Gorowitz and B. Harned, "Measurements of Velocity and Momentum with a Pulsed T-Tube Plasma Generator," Missile and Space Vehicle Dept., GE Co., Philadelphia, Pa., Tech. Information Series No. R59SD314; February 20, 1959.

[5] R. Weinstock, "Calculus of Variations with Applications to Physics and Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., 1st ed.; 1952.

[6] R. Courant and D. Hilbert, "Methods of Mathematical Physics," Interscience Publ., Inc., New York, N. Y.; 1953.

[7] P. Mostov, J. Neuringer, and D. Rigney, "Electromagnetic Acceleration of a Plasma Slug," Plasma Propulsion Lab., Republic Aviation Corp., Farmingdale, N. Y., PPL-TR-60-5; February 15, 1960.

[8] "Plasma Acceleration," S. Kash, Ed., Stanford University Press, Stanford, Calif., p. 88; 1960.

A Gas-Discharge Microwave Power Coupler*

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Summary—**A simple gas-discharge microwave power coupler consists of a cavity supporting two resonant modes at the same frequency, a gas discharge symmetrically located in the cavity with respect to each mode, and a magnetic field arranged to be perpendicular mutually to the electric fields of both modes in the region of the discharge. Power is coupled between modes by the interaction of the electrons with the electric fields of the modes in the presence of a magnetic field. If power is coupled into one mode with a suitably oriented probe, power can be coupled out of the other mode to a load by another probe. The degree of coupling is a direct function of the magnetic field and the electron density in the discharge, which may be controlled by means of an external circuit governing the current in the discharge. A large gyromagnetic resonance effect is observed when $B \cong \omega m/e$ and this resonance broadens for increasing electron density. When either the magnetic field is zero or the discharge is extinguished, no power is transferred. The mechanism has possible microwave applications for controlled or automatic power switching, amplitude modulation of power, power division, and sensing of either electron density or magnetic-field strength for control or measurement purposes. Devices having externally controlled discharges are limited to low power of the order of milliwatts. Higher-power devices may utilize self-excited discharges. Experimental results and a mathematical theory of the coupler operation are presented after a discussion of the mechanism.**

I. Introduction

THE interaction of electron beams in longitudinal magnetic fields with transverse microwave electric fields has been used by several researchers to modulate microwave power[1–3] and is currently of interest in parametric-beam amplifiers.[4] A corresponding mechanism utilizing a low-pressure gas discharge was suggested by the group working under W. C. Allis and S. C. Brown.[5] A coupler of simple geometry has been built to demonstrate the mechanism and to provide a suitable device in which the theory and experiment can be compared.

The coupler used in this study (Fig. 1) is a cylindrical cavity supporting two TE_{111} degenerate modes which are polarized at right angles to each other and whose common resonant frequency is near 3000 Mc. A gas-discharge tube with controlled dc excitation is located coaxially through the cavity. The discharge plasma contains a high concentration of free electrons which interact with the microwave electric fields in the cavity. The remaining essential element is a uniform axial magnetic field which pervades the region of the discharge and has uniform flux densities up to 0.15 weber per square meter (Fig. 2).

A discriminating coupling probe excites only one of the two resonant modes. Another similar coupling device, located at 90 degrees with respect to the first, couples the other mode to a load. Practically no power is transferred between modes if the discharge or magnetic field is absent. If the discharge and the magnetic field are present, however, there can be considerable interaction between modes. Power is then coupled from the generator, through the coupler, to the load. The amount of coupling can be controlled at will by varying either the magnetic-field strength or the electron concentration.

The electron concentration is independently controlled by the dc current exciting the discharge tube, unless the RF power is strong enough to cause a noticeable increase in the degree of ionization. The RF enhanced case is not considered in this study, and this limits to a few milliwatts, the power levels to be considered. The present concept can be extended to the high-power case by observing that the electron concentration is a function of the power and magnetic field, and is no longer an independent variable.

Before the coupler is analyzed the interaction medium and mechanism will be described.

II. The Interaction Medium and Mechanism

The electrons taking part in this interaction are in the plasma of a low-pressure electrical gas discharge. A plasma consists of approximately equal concentrations of electrons and positive ions, with the concentration in a cylindrical tube decreasing from a maximum value at the center to zero at the walls. The axial variation is negligible. Static electric fields within the discharge are low and will be ignored. The ions are slightly above room temperature and are too massive to interact appreciably with the RF electric fields, while the electrons react readily.

New plasma electrons are constantly being created within the discharge by the ionizing collisions between atoms and electrons which are at a high temperature (in the tens of thousands of degrees). The electrons are lost at the same rate, primarily by diffusion to the walls where they recombine with positive ions. As they move about within the plasma, they have many collisions

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† Electronic Tube Div., Sperry Gyroscope Co., Great Neck, L. I., N. Y.

[1] J. S. Donal and R. R. Bush, "A spiral-beam method for the amplitude modulation of magnetrons," Proc. IRE, vol. 37, pp. 375–382; March, 1950.

[2] C. L. Cuccia and J. S. Donal, Jr., "The electron coupler, a spiral-beam UHF modulator," *Electronics*, vol. 23, pp. 80–85; March, 1950.

[3] F. K. Willenbrock and S. P. Cooke, "Interaction of a spiral electron beam and a resonant microwave cavity," *J. Appl. Phys.*, vol. 21, pp. 114–125; February, 1950.

[4] R. Adler, G. Hrbek, and G. Wade, "The quadrupole amplifier, a low-noise parametric device," Proc. IRE, vol. 47, pp. 1713–1723; October, 1959.

[5] S. C. Brown, *et al.*, "Quarterly Progress Report of Research Laboratory of Electronics," Mass. Inst. Tech., Cambridge, Mass., p. 51; January, 1947.

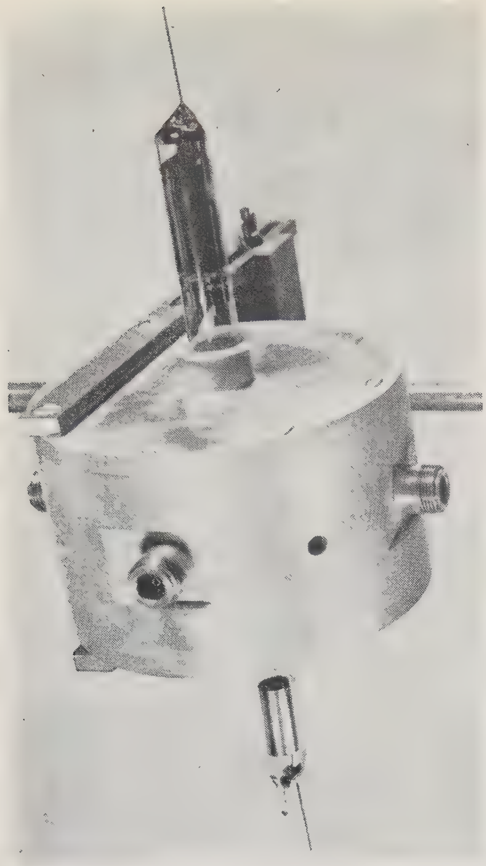


Fig. 1—An electron gas inter-mode coupler with discharge tube in place.

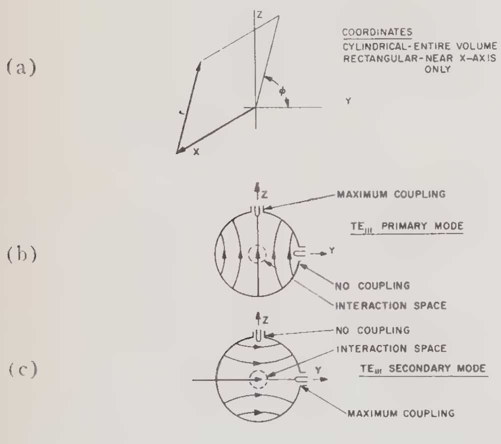


Fig. 2—The coordinate systems and the electric field configurations.

with gas molecules and occasionally with ions or other electrons, at which time they are scattered in such random fashion that they can be considered to have no memory of the direction from which they came. These collisions occur at an average frequency v. If the period between collisions is long compared to the period of the RF fields, an electron will oscillate several cycles between collisions as it interacts with the fields, and efficient interaction will take place. If the mean collision time is short, the interaction mechanism will be very lossy, as the RF energy absorbed by the electron is soon randomized and "heats" the electron gas. The analysis and discussion considers the "free electron" case primarily, with the effect of collisions being shown experimentally. The presence of an axial magnetic field causes no differences in these plasma concepts that are significant to the operation of the device, although quantitative changes will be observed.

The general equation for the motion of an electron in superimposed electric and magnetic fields is:

$$\ddot{\boldsymbol{r}} - \frac{e}{m} \boldsymbol{B} \times \dot{\boldsymbol{r}} = -\frac{e}{m} \boldsymbol{E} \qquad (1)$$

where $\boldsymbol{r}$ is the position vector and the dots represent differentiation with respect to time. If it is assumed that $\boldsymbol{B}$ is uniform in space and constant in time (or practically so within the excursion of an electron during an RF cycle) and the electric fields are time varying only, this is a linear nonhomogeneous equation with constant coefficients. This shows that the trajectory consists of the sum of the motions caused by each additive term of which $\boldsymbol{E}$ may be composed, including the case where $\boldsymbol{E}=0$. The latter case represents the motion without electric fields, which shall be called drift motion. This motion is described by a helix whose axis is in the direction of B and whose angular frequency is the cyclotron frequency, ω_g. If $\boldsymbol{E}$ is further assumed to consist of two mutually perpendicular RF fields having the same frequency but any phase and amplitude, and which are also perpendicular to $\boldsymbol{B}$, the combined motion related to these two RF fields will be elliptical in the plane of the RF fields. The complete motion is then an ellipse of frequency ω, superimposed upon a circle of frequency ω_g, all in the plane of the electric fields and moving at a constant velocity in the direction of the magnetic field.

On the average there is no net power transferred between the RF fields and the circular motion because the force and the motion are not at the same frequency. Net power transfer can occur only with the elliptical motion; therefore, only the elliptical motion need be considered to obtain the coupling characteristics. A qualitative picture of the coupling phenomenon is obtained as follows.

Consider the forced motion of an electron (and ignore the drift component) in a plane-polarized RF electric field. The motion will be an oscillation back and forth in the direction of the electric field. Impose a constant uniform magnetic field perpendicular to the electric field. As the electron oscillates, there is a force perpendicular to the direction of motion, and this force produces an elliptical motion. There is now a component of cyclic motion in a direction mutually perpendicular to both fields, and this motion of a charged particle constitutes a current that excites a second electric RF field orthogonal to the first. If there is a means of absorbing or radiating power from the second field, power will be transferred from the first via the electron to the second, and hence, to a load. In an electron gas in uniform

fields, the driven motions of all the electrons are synchronous in phase and of equal amplitude, so that the total electronic charge constitutes an oscillating space-charge current. A simple coupler utilizing this mechanism will now be analyzed.

III. Analysis

The objectives of the analysis are to determine the input admittance of the coupler as a function of electron concentration, magnetic field strength, load, and fixed properties of the cavity, and to evaluate the power delivered to the load as a function of the same variables. These objectives are achieved by analyzing the electronic space-charge current caused by the electric fields of the two modes in the cavity and interpreting these fields and currents in terms of a lumped-constant equivalent circuit. The analysis begins with the RF electric fields.

A. The RF Fields

The electric fields are pictured in Fig. 2 along with the applicable cylindrical and rectangular coordinate systems. The primary mode, which is driven by the generator, is polarized in the Z direction; and the secondary mode, which feeds the load, is polarized in the Y direction. Expressions for the TE_{III} fields are generally given in cylindrical coordinates and are well known.[6] In the interaction space close to the axis, they can be approximated for the two modes with Z and Y polarization, respectively, by the simple rectangular forms:

$$E_Z \cong E_{Z_0} \sin\left(\frac{\pi x}{L}\right) \exp(j\omega t) \tag{2}$$

and

$$E_Y \cong E_{Y_0} \sin\left(\frac{\pi x}{L}\right) \exp(j\omega t) \tag{3}$$

where E_{Z_0} and E_{Y_0} are respectively the complex Z-directed and Y-directed RF electric fields at the center of the cavity, including magnitude and phase information. The cavity of length L is one-half wavelength long, and it is excited at a circular frequency ω.

B. The Electronic Motion

Let $\boldsymbol{r}=x\boldsymbol{i}+y\boldsymbol{j}+z\boldsymbol{k}$ where $\boldsymbol{i}$, $\boldsymbol{j}$, and $\boldsymbol{k}$ are unit vectors in the x, y, and z directions, respectively. Also, let $\boldsymbol{B}=B_x\boldsymbol{i}$. $\boldsymbol{E}$ will be represented by $E_Y\boldsymbol{j}+E_Z\boldsymbol{k}$ from (2) and (3). The equation of motion (1) then becomes

$$\ddot{x} = 0 \tag{4}$$

$$\ddot{y} + \left(B_x \frac{e}{m}\right)\dot{z} = -\frac{e}{m} E_{Y_0} \sin\left(\frac{\pi x}{L}\right) \exp(j\omega t) \tag{5}$$

$$\ddot{z} - \left(B_x \frac{e}{m}\right)\dot{y} = -\frac{e}{m} E_{Z_0} \sin\left(\frac{\pi x}{L}\right) \exp(j\omega t). \tag{6}$$

[6] C. G. Montgomery, R. H. Dicke, and E. M. Percell, "Principles of Microwave Circuits," Mass. Inst. Tech. Rad. Lab. Ser. No. 8, McGraw-Hill Book Co., Inc., New York, N. Y.; 1948.

Eq. (4) indicates that the x component of motion does not contribute to the coupling; therefore, the x component will be omitted from the remainder of the discussion.

Harmonic driving forces result in harmonic responses in linear systems, so that time can be removed from the remaining two equations, (5) and (6), by the substitutions $z=Z_0 \exp(j\omega t)$ and $y=Y_0 \exp(j\omega t)$, yielding a pair of simultaneous algebraic equations which can be solved to give

$$Z_0 = \frac{e}{\omega^2 m} \frac{E_{Z_0} - jbE_{Y_0}}{(1-b^2)} \sin\left(\frac{\pi x}{L}\right) \tag{7}$$

and

$$Y_0 = \frac{e}{\omega^2 m} \frac{E_{Y_0} + jbE_{Z_0}}{(1-b^2)} \sin\left(\frac{\pi x}{L}\right). \tag{8}$$

Here, b is the ratio of the existing magnetic field to the value necessary to produce cyclotron resonance at frequency ω; thus

$$b = \frac{B_x e}{\omega m}. \tag{9}$$

It is apparent from (7) and (8) that as b approaches unity from either direction, the amplitude of the motion becomes very large. If the system is dissipative, however, the amplitude does not go to infinity because the numerators also become zero and the amplitude remains finite.

C. The Equivalent Circuit

The operation of the coupler can be described in terms of the equivalent circuit shown in Fig. 3. The two resonant modes are represented by two parallel res-

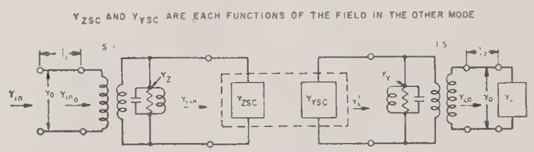


Fig. 3—Complete equivalent circuit of the electron gas inter-mode coupler.

onant circuits having admittances Y_Z and Y_Y. The load and generator are coupled to their respective modes by ideal transformers of turns ratio S:1 and two lengths of line l_1 and l_2 which are normally considered as part of the transmission lines connecting the coupler to the generator and the load, respectively. Electronic interaction, which is the only connection between the modes, is represented by the equivalent space-charge admittances Y_{ZSC} and Y_{YSC}.

A quantitative analysis of the above circuit requires that the various parameters be defined in a self-consistent manner. This is done on the basis of energy and the electric fields at the center of the cavity. Using these

definitions, it is possible to measure or calculate all of the elements in the equivalent circuit. First, two "voltages" are defined:

$$V_Z = -dE_{Z_0}\exp(j\omega t) = V_{Z_0}\exp(j\omega t) \tag{10}$$

$$V_Y = -dE_{Y_0}\exp(j\omega t) = V_{Y_0}\exp(j\omega t) \tag{11}$$

where V_{Z_0} and V_{Y_0} are respectively the complex Z-directed and Y-directed voltages, including magnitude and phase information, and d is the cavity diameter.

The space-charge current can be expressed as

$$J(x, y, z, t) = -eN(x, y, z)\mathbf{V} = -Ne\frac{\partial r}{\partial t}$$
$$= -eNj\omega(y\mathbf{j} + z\mathbf{k}) \tag{12}$$

where $N(x, y, z)$ is the electron concentration at any point in the discharge. If one defines

$$n(x, y, z) = \frac{N(x, y, z)}{N_L} = \frac{N(x, y, z)}{\left(\frac{\omega^2\epsilon_0 m}{e^2}\right)}, \tag{13}$$

which is the relative value of N to the value for Langmuir space-charge resonance, the current components can be written

$$J_Z(x, y, z, t) = -\omega\epsilon_0 n\frac{(bE_Y + jE_Z)}{(1-b^2)} \tag{14}$$

and

$$J_Y(x, y, z, t) = \omega\epsilon_0 n\frac{(bE_Z - jE_Y)}{(1-b^2)}. \tag{15}$$

These expressions are useful in computing the interaction impedances Y_{ZSC} and Y_{YSC}. The concept of real and susceptive (reactive) power is used to define Y_{ZSC} and Y_{YSC} in terms of the fields and current densities as follows:

$$P_{ZR} + jP_{ZI} = \frac{1}{2}(G_{ZSC} + jB_{ZSC})|V_Z|^2$$
$$= \frac{1}{2}\int_{\tau_j}(E_Z^*\cdot J_Z)d\tau, \tag{16}$$

and

$$P_{YR} + jP_{YI} = \frac{1}{2}(G_{YSC} + jB_{YSC})|V_Y|^2$$
$$= \frac{1}{2}\int_{\tau_j}(E_Y^*\cdot J_Y)d\tau, \tag{17}$$

where τ_j is the volume of the discharge in the cavity. If the integrations implied in (16) and (17) are carried out and the ratio

$$\frac{E_{Z_0}}{E_{Y_0}} = p + jq \tag{18}$$

is defined, the real and imaginary components of the interaction admittances become:

$$G_{ZSC} = -\frac{\omega\epsilon_0\tau_j}{2d^2}n_a\frac{b}{(1-b^2)}\cdot\frac{p}{(p^2+q^2)}, \tag{19}$$

$$B_{ZSC} = \frac{\omega\epsilon_0\tau_j}{2d^2}n_a\frac{\left(\frac{bq}{p^2+q^2}\right)-1}{(1-b^2)}, \tag{20}$$

$$G_{YSC} = \frac{\omega\epsilon_0\tau_j}{2d^2}n_a\frac{bp}{(1-b^2)}, \tag{21}$$

and

$$B_{YSC} = \frac{\omega\epsilon_0\tau_j}{2d^2}n_a\frac{bq-1}{(1-b^2)}. \tag{22}$$

The term n_a is the average relative electron concentration, defined as

$$n_a = \frac{1}{\tau_j}\int_{\tau_j} n\,d\tau. \tag{23}$$

The shunt conductances G_Z and G_Y of the resonant modes without space-charge loading will be defined to give the correct dissipation when the voltages V_Z and V_Y are applied. The usual expression for the unloaded Q of a cavity mode yields:

$$G_Z = G_Y = \frac{\omega C}{Q_0}. \tag{24}$$

The electromagnetic energy stored in one cavity mode can be expressed as:

$$W_c = \frac{1}{2}C|V_{Z_0}|^2 = \frac{1}{2}\epsilon_0\int_{\tau_c}(E_1\cdot E_1^*)d\tau, \tag{25}$$

where E_1 is the value of the electric field of one mode at any point in the cavity volume, τ_c. This gives

$$C = \frac{\epsilon_0\tau_0}{d^2}, \tag{26}$$

where τ_0 is defined for the TE_{III} mode by

$$|E_{Z_0}|^2\tau_0 = \int_{\tau_c}(E_1\cdot E_1^*)d\tau = 0.238\tau_C|E_{Z_0}|^2. \tag{27}$$

C can now be computed, and Q_0 is easily measured experimentally with the discharge extinguished. Evaluation of the remaining constants S and Y_L is conveniently done experimentally. The explanation will be omitted as the methods are not unusual and the details are not essential.

D. Circuit Relationships

It is convenient to define Y_L' as the load seen by the electrons, *i.e.*, $Y_Y + S^2 Y_{L_0}$ (Fig. 3). As can be shown for any circuit which is cut through a single loop so as to

isolate the two halves of the circuit,

$$Y_{YSC} + Y_L' = 0, \tag{28}$$

where Y_{YSC} must have the characteristics of a generator, *i.e.*, negative conductance. Eqs. (21), (22), and (28) can be combined to give expressions for p and q that determine the relative amplitude and phase of the RF fields in the two modes

$$p = -\frac{\Gamma'}{n_a}\cdot\frac{1-b^2}{b} \tag{29}$$

and

$$q = -\frac{\Lambda'}{n_a}\cdot\frac{(1-b^2)}{b} + \frac{1}{b}, \tag{30}$$

where

$$\Gamma' = \frac{2d^2 G_L'}{\omega\epsilon_0\tau_j}, \tag{31}$$

and

$$\Lambda' = \frac{2d^2 B_L'}{\omega\epsilon_0\tau_j}. \tag{32}$$

The dimensionless load parameters Γ' and Λ' are constants at a given frequency but are frequency-sensitive because of the parallel resonant circuit and the possible frequency sensitivity of the load.

It is now convenient to define an electronic admittance function,

$$\eta = \alpha + j\beta = \frac{2d^2}{\omega\epsilon_0\tau_j} Y_{ZSC} = \frac{2d^2}{\omega\epsilon_0\tau_j}(G_{ZSC} + jB_{ZSC}). \tag{33}$$

If (29) and (30) are substituted into (19) and (20), (33) becomes

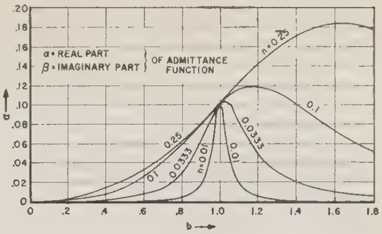


Fig. 4—Real part of electronic admittance function vs relative magnetic flux density for different values of realtive electron concentration, all at $\Gamma'=0.1$ and $\Lambda'=0.0$.

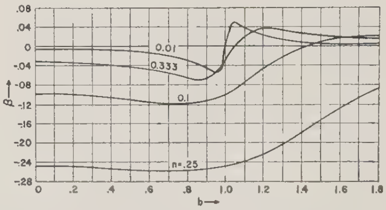


Fig. 5—Imaginary part of electronic admittance function vs relative magnetic flux density for different values of relative electron concentration, all at $\Gamma'=0.1$ and $\Lambda'=0.0$.

since b always appears as the square in (34) and (35) that this coupler exhibits no nonreciprocal properties in either phase or amplitude.

The input admittance expressed at the input terminals of the input transformer is simply:

$$Y_{in_0} = \frac{1}{S^2}(Y_Z + Y_{ZSC}). \tag{36}$$

Since Y_Z will usually be small compared to Y_{ZSC}, especially near resonance, the input admittance to the coupler will look very much like η, from which it is readily computed. It is obvious that all of the steady-state properties of the coupler as a circuit element

$$\alpha = \frac{b^2\Gamma'}{\left[\left(\frac{\Gamma'}{n_a}\right)^2 + \left(\frac{\Lambda'}{n_a}\right)^2\right](1-b^2)^2 - 2\left(\frac{\Lambda'}{n_a}\right)(1-b^2) - 1}, \tag{34}$$

and

$$\beta = -n_a\left\{\frac{\left[\left(\frac{\Gamma'}{n_a}\right)^2 + \left(\frac{\Lambda'}{n_a}\right)^2 - \left(\frac{\Lambda'}{n_a}\right)\right](1-b^2) - \left(\frac{\Lambda'}{n_a}\right) + 1}{\left[\left(\frac{\Gamma'}{n_a}\right)^2 + \left(\frac{\Lambda'}{n_a}\right)^2\right](1-b^2)^2 - 2\left(\frac{\Lambda'}{n_a}\right)(1-b^2) + 1}\right\}. \tag{35}$$

These functions are plotted in Figs. 4 and 5 for the case where $\Gamma'=0.1$ and $\Lambda'=0$. Curves are plotted for various values of n_a as a function of b. Gyromagnetic resonance is observed to occur in the vicinity of $b=1$. Increasing the electron concentration causes a broadening of the resonance and a shift of the maximum of the real part of the electronic admittance function toward higher values of b. β shows a typical resonance characteristic especially at lower electron concentrations. It is apparent (within the limits of certain assumptions) can be obtained from the equivalent circuit. For example, the power to the load can be computed.

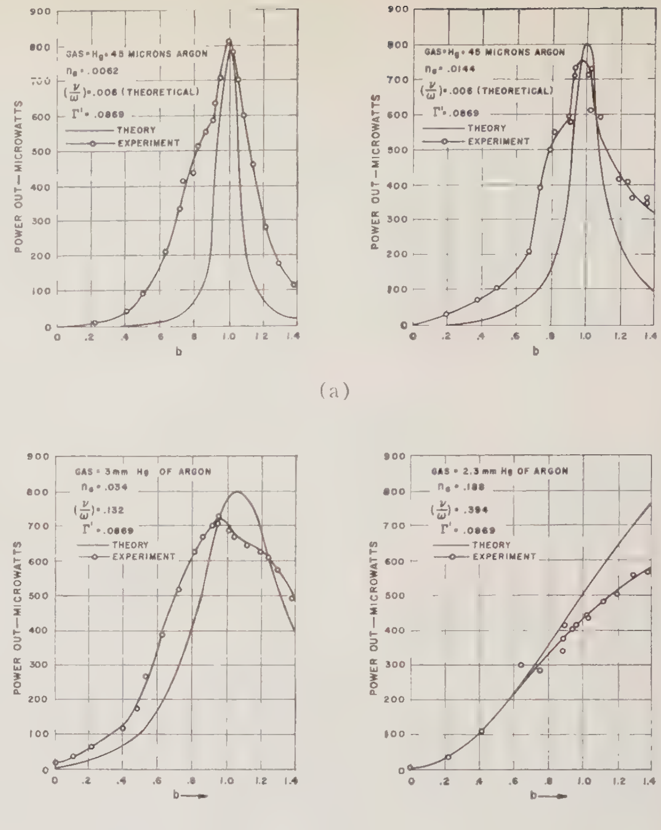
E. Power to the Load

In the experiment, the coupler is driven by a generator through a long length of lossy line. This is equivalent to an ideal constant-current generator of current I_g shunted by the characteristic admittance of the line

Y_0. The load in the experiment is a thermistor in a mount, having a low standing-wave ratio, so that $Y_L = Y_0$, which gives $\Lambda' = 0$. The cavity modes are in resonance, so that $Y_Z = Y_Y = G_Z$. The power will be computed for this case. The ideal generator sees input admittance Y_{in_0}. Since the plasma losses are neglected, the power delivered to Y_L' is the same as that delivered to Y_{ZSC}. The power delivered to the load is

$$P_L = P_L' \frac{G_L' - G_Z}{G_L'} = P_L'\left(1 - \frac{Q_L'}{Q_0}\right)$$
$$= \frac{|I_g|^2}{|Y_0 + Y_{in_0}|^2}\left(\frac{\omega\epsilon_0\tau_j}{2d^2S^2}\right)\left(1 - \frac{Q_L'}{Q_0}\right)\alpha, \qquad (37)$$

where Q_L' is the Q of the cavity loaded by $G_Y + S^2Y_0$. Several theoretical and experimental curves showing the variation of power to the load with n_a and b are shown in Fig. 6. The experimental results shown in these figures will be discussed in the next section.



(a)

(b)

Fig. 6—Experimental and theoretical curves of power out for a constant incident power of one milliwatt.

IV. Experimental Results

A. Objectives and Conditions of the Experiments

The experiments emphasize measurement of the dependence of the intermode power-transfer characteristics upon the electron concentration and the magnetic flux density. The theoretical expression for transferred power is determined in the previous section in terms of these same variables and other measurable circuit parameters.

The reported experimental work was done at 3000 Mc, but the theory is generally applicable at other frequencies, provided $(\nu/\omega) \ll 1$. It is assumed that high-frequency power lost to the gas is not large enough to increase the electron concentration to a noticeable degree. The theory is still applicable at higher power levels but n_a would no longer be an independent variable and n would be a function of distance along the axis. The collision-frequency and power-level requirements were met at 3000 Mc by using gas pressures less than 3 mm Hg and RF powers less than 2 milliwatts.

B. Measurement of n_a and (ν/ω)

No adequate theory was available for the measurement of n_a and (ν/ω) in the presence of a strong magnetic field at the time of the experiments. Since then, the work of Bickerton and von Engel[7] indicates that satisfactory probe measurements of n can be made. It was decided, therefore, to measure n_a and (ν/ω) without the magnetic field and to use these values in the computations, bearing in mind that they are not entirely correct. The work of Bickerton and von Engel indicates that these measurements may be low by a factor of three or four at high values of magnetic field. If the electron concentration near cyclotron resonance is computed from the measured half-power bandwidth and the coupler theory, the values of n_a as measured at $b=0$ are indicated to be low by factors of $3\frac{2}{3}$, 2 and $1\frac{3}{4}$ respectively near $b=1$ for the first three cases shown in Fig. 6(a) and 6(b).

The measurements at $b=0$ utilize a microwave method based on the work of Slater (Eq. III 82)[8] and Bailey (Eq. 6).[9] This permits n_a and (ν/ω) to be measured by observing the shift of the resonant frequency and the change in the reciprocal of the unloaded Q of a cavity when the discharge is initiated. For brevity, the derivation is omitted (see Olthuis[10] for complete details). The resulting expressions for loss factor and average relative electron concentration are

$$\frac{\nu}{\omega} = \frac{1}{2\delta Q_j} \qquad (38)$$

and

$$n_a = \frac{4\tau_0}{\tau_j}\left[\left(\frac{1}{2\delta Q_j}\right) + 1\right]\delta, \qquad (39)$$

[7] R. J. Bickerton and A. von Engel, "The positive column in a longitudinal magnetic field," *Proc. Phys. Soc.* (London), vol. 69B, pp. 468–481; April, 1956.

[8] J. C. Stater, "Microwave electronics," *Revs. Mod. Phys.*, vol. 18, pp. 441–512; October, 1946.

[9] V. A. Bailey, "Electro-magneto-ionic optics," *J. Proc. Roy. Soc. NSW*, vol. 82, pp. 107–113; June, 1948.

[10] R. W. Olthuis, "A study of the coupling induced between two modes in a microwave cavity by a gas discharge in a magnetic field," *Microfilm Abst.*, vol. 11, no. 2, Publ. No. 2425, pp. 106–113; 1951.

where

$$\delta = \frac{f_{\text{final}} - f_{\text{initial}}}{f_{\text{initial}}} \tag{40}$$

and

$$\frac{1}{Q_j} = \frac{1}{Q_{0_{\text{final}}}} - \frac{1}{Q_{0_{\text{initial}}}}. \tag{41}$$

Curves of n_a which were obtained by this method using one of the T_{III} modes in the coupler cavity are shown in Fig. 7. The electron temperature and, therefore, (ν/ω) are relatively independent of the tube current, so only one measurement of (ν/ω) is reported for each tube.

The three tubes mentioned in Fig. 7 are the sealed-off gas-filled tubes with which all the reported coupler data were obtained. The first contains mercury vapor plus 45 microns Hg of argon; the second, 0.3 mm Hg of argon; and the third, 2.3 mm Hg of argon. The mercury tube has a hot cathode; the others have cold nickel cathodes and anodes. The tubes are one-half inch in diameter, and the electrodes are six inches apart and are located outside the cavity.

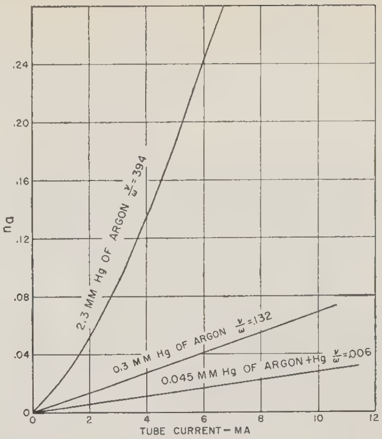


Fig. 7—Average relative electron concentration vs tube current.

C. Test Circuit

A schematic diagram of the experimental setup used to measure the characteristics of the coupler is shown in Fig. 8. A reflex klystron generator is square-wave amplitude modulated. The output is fed to a slotted section by a 50-ohm lossy coaxial cable having 8.3-db attenuation. More isolation would have been desirable, but the power levels became prohibitively low for measurement with available equipment. A constant generator frequency is maintained during an experiment by reference to the wavemeter.

The slotted section, probe, and crystal serve for measurement of both the standing-wave ratio and the input power. The tuned audio amplifier includes an output meter together with a range switch which, when combined with the decade attenuator, enables the input power level and standing-wave ratio to be measured.

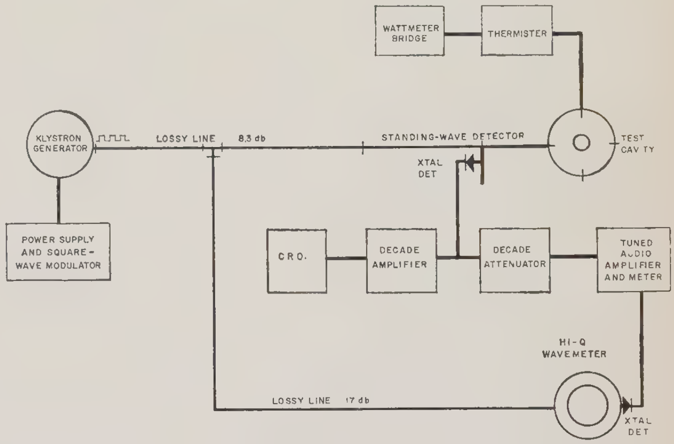


Fig. 8—Experimental setup.

D. Performance Characteristics

A representative set of curves showing the coupler performance as a function of the magnetic flux density is presented in Fig. 9. The shift of the position of standing-wave minimum resembles the electronic susceptance curve (Fig. 5), and the output-power curve is similar to the electronic conductance curves (Fig. 4).

One anomoly that can be noted in the qualitative comparison is the ripple in the output-power curve just before the resonance peak. The ripple has the appearance of a small, sharp resonance superimposed on the broader anticipated resonance curve. This occurs in all measurements, but is particularly noticeable for low electron concentrations. No explanation is offered at this time.

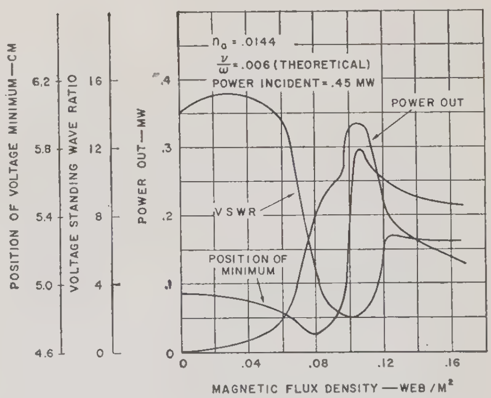


Fig. 9—Typical performance curves.

It can be observed in Fig. 6 that the experimental resonances are broader than the corresponding theoretical resonances. The amplitudes of the peaks, however, are in close agreement. Since the amplitude of the peak is theoretically relatively independent of n_a, these data indicate that the measurements of the circuit parameters other than n_a are in good agreement, while n_a (which is

an important factor in determining the breadth of the resonance) is undervalued. The magnetic field was also slightly barrel shaped, which tends to broaden the resonance curve.

The last three curves of Fig. 6 show the effect of increasing loss factor due to collision losses. The experimental amplitude falls below the lossless theoretical curves more and more as (ν/ω) increases. The curve for $n_a=0.188$ represents a condition for which an electron makes about two collisions per cycle. The maximum over-all efficiency is over 90 per cent for the first case and decreases successively for the others.

Effects of spurious ionic plasma oscillations were detected under some conditions. When making the (ν/ω) and n_a measurements for the tube filled with 2.3 mm Hg of argon, a modulation of the position of minimum in the slotted line was detected, and this limited the range of discharge-tube current over which measurements could be made to less than 8 ma. The data presented in Fig. 6 for this tube were taken at a discharge current of 5 ma.

Fig. 10 is an experimentally determined plot of power to the load, showing a family of resonance curves for different values of arc current vs magnetic-field strength for a constant value of incident power. The relative average electron concentration increases from 0.0035 to 0.014 as the discharge current increases from 1 to 5 ma. Gyromagnetic resonance is computed to occur at $B=0.106$ webers per square meter. The resonance maximum is seen to occur approximately at this value and to move toward the right as n_a increases in analogous fashion to α in Fig. 4. The wiggle in the curve just before resonance is clearly evident.

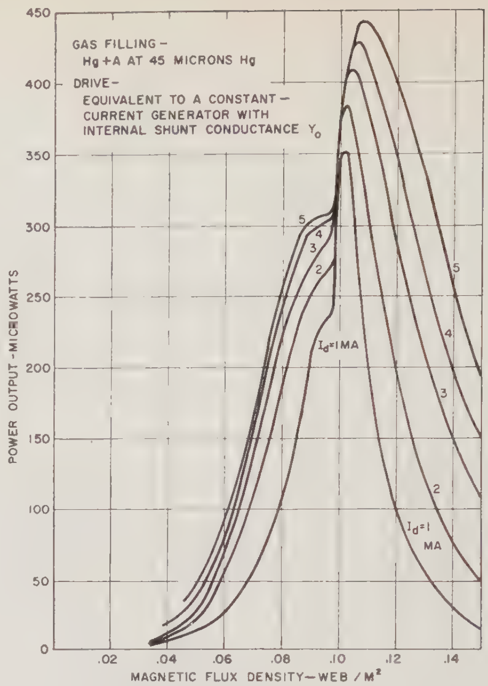


Fig. 10—Power output vs magnetic flux density with discharge current as the parameter.

V. Discussion

The last figure is particularly interesting because it indicates the way the device can be used to amplitude modulate microwave power. As an on-off device, either the discharge current or the magnetic field can be modulated from zero to an appropriate fixed value. For linear modulation, either B can be modulated in the range 0.105 to 0.13 at fixed values of I_b, or I_b can be modulated for a fixed value of B in the same range. Practical use of this device as a linear modulator of microwave signals at audio or video frequency is limited by the inability of the electron concentration in the discharge to follow the modulating signal rapidly, and the difficulty of modulating a magnet of this size. This coupler exhibits no nonreciprocal properties.

A converse application of the mechanism suggests a device for sensing the magnitude of I_b, N or B for control or measurement purposes. Practical applications for the coupling mechanism may be found in the higher power range where the electron concentration is no longer controlled by the power supplied from an external source, but rather is subject to the RF power itself. The possibility of automatic switching devices based on power level is suggested. For instance, if there is no external power supply for the discharge, or the applied potential is maintained below the breakdown potential, low-power RF energy will be reflected by the coupler. When the RF power reaches a critical amplitude, the discharge will break down, and with a proper value of magnetic field, the discharge will be efficiently coupled to the load.

VI. Acknowledgment

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Correspondence

Proposal for a Pulsed Ferromagnetic Microwave Generator*

The generation of microwave radiation by means of a system of spins was accomplished several years ago with the invention of the paramagnetic maser. It has been realized for some time that an analogous ferromagnetic device would have two major potential advantages: it could provide much larger amounts of power because of the larger number of spins available and because nearly all the spins rather than a fraction exp $(-\hbar\omega/kT)$ are lined up; and it could operate at room temperature.

Several schemes have been proposed and utilized[1-4] for such a device. In all of these, the magnetization of a saturated ferromagnet is transiently brought into a state in which it is not aligned with the magnetic field and is then made to radiate into a circuit at the free precession frequency. We are here proposing a new method for producing a nonaligned magnetization which may have some advantages.

For certain ranges of value of the crystalline anisotropy constants (which can be obtained in several known compounds of both cubic and hexagonal structure), the uniform magnetization of a saturated monocrystalline ellipsoid can assume two or more orientations. The energy is lowest in one of these orientations; the others are metastable, corresponding to local but not absolute minima of the energy surface.

The sample can be put into a metastable state by turning on a field large enough to produce an absolute energy minimum in the direction of the field. The field can then be reduced or rotated until some other direction of the magnetization becomes absolutely stable (corresponding to a deeper energy minimum). If now the field is changed further until the metastable minimum disappears, the magnetization will precess about the direction of the new (absolute) minimum and can radiate into a microwave circuit.

An example of such a process is furnished by a disc of cubic material with positive anisotropy cut parallel to a (100) plane, and thus containing four easy (100) directions. Let the sample be saturated by applying a field larger than the anisotropy field H_A along one of these directions. If the applied field H is now reduced to some value in the range $\frac{1}{4}H_A<H<H_A$, and rotated in the plane of the disc, the magnetization becomes metastable as soon as the angle of rotation exceeds 45°. At some angle of rotation between 45° and 90°, the metastable direction becomes unstable, and the magnetization precesses about a new direction. The conditions for the instabilities and equilibria are readily computed. We can get an idea of the magnitudes involved from a numerical example; if $H=\frac{1}{2}H_A$, the instability first occurs when the angle of rotation is 55.7° and the initial departure of the magnetization from its new equilibrium orientation is 48.7°.

The radiated frequency depends on the values of H, the first- and second-order anisotropy constants K_1 and K_2, and the magnetization M, and can be computed readily. For a numerical example consider a (100) disc with a longitudinal demagnetizing factor of 0.8 (transverse demagnetizing factor 0.1) cut from a material with $4\pi M=5000$ gauss, first-order anisotropy field $H_A=2000$ oersteds, and $g=2.3$. The radiated frequency varies from 8 to 12 kMc as the applied field is changed from 1700 to 1100 oersteds. The energy that is potentially available for the radiation is the difference between the metastable and final energy minima which, for this example, varies from 5×10^4 to 10^4 ergs per cubic cm over this frequency range. Assuming a 0.01-cm³ sample, a 1-μsec pulse, and an efficiency of 10 per cent, this would furnish a pulse power of 1 to 5 watts.

The constants of the material for this numerical example have been chosen from the values characteristic of dilute (approximately 20 per cent) cobalt ferrite. The main difficulty in using this substance in such an experiment is the rather large linewidth (low intrinsic Q), which implies a restriction to extremely short pulses. For a first experiment it may be easier to use the anomalous anisotropy observed in yttrium iron garnet at liquid helium temperatures.[5]

One phenomenon associated with the proposed experiment should be mentioned. It has been pointed out by Schaug-Pettersen[6] that under certain conditions a class of spinwaves in a ferromagnetic specimen can exhibit unstable growth. In the experiment proposed here, the instability of the uniform precession which we wish to exploit is always preceded by such a spinwave instability. Since the growth of spinwaves does not contribute to the microwave radiation,[6] it is necessary to traverse the spinwave instability region rapidly. Thus the applied field cannot be rotated slowly, but must be pulsed over a portion of the cycle. With the disc geometry described here the spinwave instability region can be made quite narrow and the growth rate can be kept relatively small. Exact calculation is difficult, but a rough estimate indicates that the requirements are well within the capabilities of present pulse techniques.

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* Received by the IRE, March 1, 1961.

[1] R. V. Pound, "Microwave Pulse Generator," U. S. Patent No. 2,873,370; February 10, 1959.

[2] F. R. Morgenthaler, "Microwave radiation from ferrimagnetically coupled electrons in transient magnetic fields," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 6–11; January, 1959.

[3] B. J. Elliott, T. Schaug-Pettersen and H. J. Shaw, "Pulsed ferrimagnetic microwave generator," *J. Appl. Phys.*, vol. 31, pp. 400S–401S; 1960.

[4] P. Wolf, *J. Appl. Phys.*; March, 1961.

[5] J. F. Dillon Jr., "Ferrimagnetic resonance in yttrium iron garnet at liquid helium temperatures," *Phys. Rev.*, vol. 111, pp. 1476–1478; September, 1958.

[6] T. Schaug-Petersen, "Growing spin waves in ferrites in unstable equilibrium," *J. Appl. Phys.*, vol. 31, pp. 382S–383S; May, 1960.

RF-Induced Negative Resistance in Junction Diodes*

Experiments in using junction diodes as harmonic generators have brought to our attention a peculiar effect in the bias circuit. It was noticed that over a region of reverse bias voltage, an increase in voltage resulted in a bias-current decrease, indicating a negative resistance. This effect existed only while the fundamental RF power was driving the diode but was present both when the power at the harmonic frequency was allowed to dissipate and when it was not.

One type of RF-induced negative resistance in junction diodes has recently been reported by Hefni.[1] However, we have observed an additional type of negative resistance in our diodes. The diodes can exhibit either of two types of negative resistance, depending on the reverse bias voltage at which the impedance of the diode is matched to the RF source. Small biases give rise to the type in which current is a multiple valued function of voltage (S type), while larger biases effect the type in which voltage is a multiple valued function of current (N type).

A circuit diagram of the experimental set-up is shown in Fig. 1. The harmonic output circuit is not shown. The RF frequency was about 350 Mc. Input power ranged from 50 to 100 mw. The oscilloscope displayed applied bias voltage on the horizontal axis and bias current on the vertical axis. The diode I-V characteristic was observed on the scope with the RF signal applied to the diode. When the matching network was correctly tuned, so that sufficient power could flow to the diode, the negative resistance characteristic appeared. Fig. 2 shows curve traces for a graded junction silicon diode (MA 4380 X) exhibiting an N-type negative resistance in Fig. 2(a) and an S-type in Fig. 2(b). It was found that both types could also be obtained with an abrupt junction diode (TIC 64).

Using the S-type negative resistance we were able to obtain astable, monostable or bistable operation of the bias circuit. In the astable mode, the period of oscillation is dictated by the time constant of the bias circuit. Output waveforms for the three modes of operation are shown in Fig. 3.

The highest frequency at which the circuit will oscillate should be limited primarily by the driving frequency because of the mechanism which is proposed to explain the negative resistance. The phenomenon of hole storage is proposed as the cause, but a current change in the bias circuit should require at least one cycle of the driving signal and may require many cycles. We are trying to develop a theory which will predict the negative resistance and establish frequency limitations.

* Received by the IRE, March 1, 1961.

[1] I. Hefni, "Effect of minority carriers on the dynamic characteristic of parametric diodes," *Electronic Engrg.*, vol. 32, pp. 226–227; April, 1960.

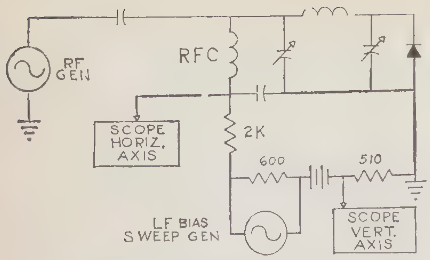
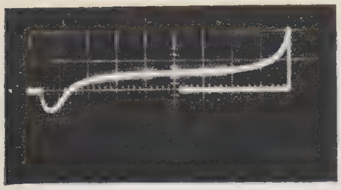
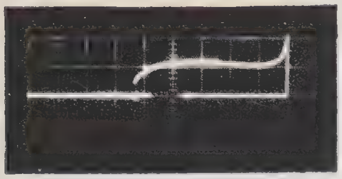


Fig. 1.—Circuit diagram.

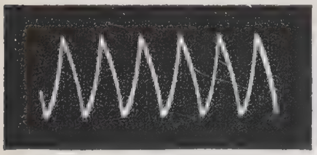


(a)



(b)

Fig. 2.—I-V Characteristics showing (a) N-type and (b) S-type negative resistance in a MA 4380 X diode. The I-V characteristic with power off is also shown. Horizontal scale: 2 v/cm, increasing to the right. Vertical scale: 0.5 ma/cm, positive upward.



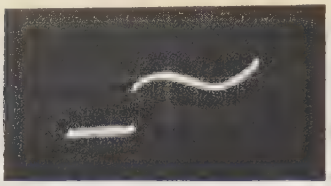
(a)

(b)

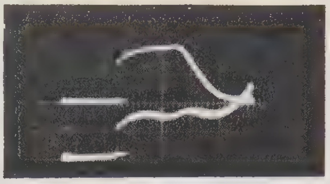
(c)

Fig. 3—Output waveforms of the bias circuit in (a) astable, (b) monostable, and (c) bistable operation. The output is about 1 v peak to peak and was taken across the 510-Ω resistor in the bias circuit. The load lines were determined by the bias voltage and the total resistance of the bias circuit. The astable frequency was 100 kc. The monostable and bistable outputs were triggered by pulses at 2 kc.

Fig. 4 exhibits two interesting features of a diode biased near zero. The first is the appearance of an additional N-type negative resistance close to zero bias. The second is that this additional negative resistance disappears when the harmonic power is drawn from the diode.



(a)



(b)

Fig. 4—(a) Negative resistance of a TI C 64 diode near zero bias. Horizontal scale: 3 v/cm centered at −6 v. Vertical scale: 0.3 ma/cm. (b) The lower curve shows cancelling of the negative resistance due to power being drawn at five times the input frequency. The upper curve shows power output vs bias voltage. Peak output is 5 mw at −5.5 v bias. Input was 50 mw at 350 Mc.

This negative resistance near zero bias is similar in appearance to an RF-induced negative resistance observed by North on welded contact diodes.[2] Capacitance variation with voltage was proposed to explain North's negative resistance. It seems reasonable that the negative resistance should disappear when harmonic power is drawn from the diode because this has the effect of adding positive resistance to the diode.

The helpful advice of I. Friedberg on this work is gratefully acknowledged.

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[2] H. C. Torrey and C. A. Whitmer, "Crystal Rectifiers," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 15, p. 401; 1948.

Note on Coherence vs Narrow-bandedness in Regenerative Oscillators, Masers, Lasers, etc.*

In several discussions with engineers and physicists engaged in pushing the art of generating coherent radiation to shorter wavelengths, I have noted that the narrow-bandedness of the radiation produced, if not confused with the coherence[1] of this radiation, is often considered to be a measure of the degree of coherence. I would like to point out that coherence is not a quantitative concept, but a qualitative one; either radiation is coherent, or it is not, regardless of bandwidth considerations. If the duration of the signal is not indefinite, due to physical limitations such as the heating of a ruby laser, or of the source of pumping radiation, either the radiation is coherent during its ephemeral duration, or it is not.

* Received by the IRE, February 8, 1961.

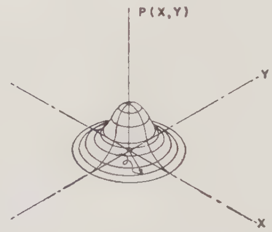
[1] The term "coherence" is used here in the sense of CW radiation which can interfere with indefinitely delayed portions of itself. It should not be confused with optical coherence, which refers to the property which two light beams have of interfering with each other when they are obtained from a common source.

Theoretical experiments, probably realizable at later times, may be useful to form a picture of the concept involved here.

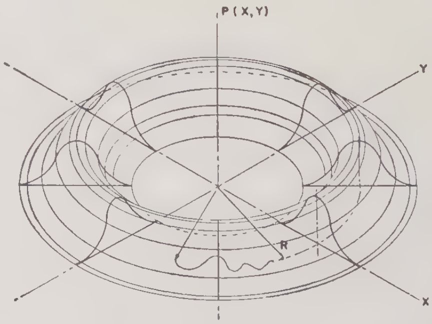
Consider a source of narrow-banded radiation which must be tested for coherence, and designate its approximate center frequency by f. If the source output is heterodyned with two assumedly available perfectly monochromatic signals proportional to cos 2π ft and sin 2π ft respectively, the two components of a phasor are obtained.

The behavior of this phasor will constitute the criterion of coherence vs noncoherence of the radiation studied.

If this radiation is incoherent, the phasor end will be observed to execute a two-dimensional random walk at a rate inversely proportional to the bandwidth of this radiation, and if the experiment is continued over a sufficiently long period of time, the statistical location of this phasor end will be describable by a probability distribution which, in most instances, will be a Gaussian "mole-hill" centered at the origin [Fig. 1(a)].



(a)



(b)

Fig. 1—Probability distribution of the phasor representing the relationship between a narrow-band signal and a perfectly monochromatic reference signal, for two cases. (a) The signal is incoherent. The distribution is a "mole-hill" centered on the origin, showing no tendency toward regulation of amplitude. (b) The signal is coherent. The "mole-run" distribution shows strong tendency to regulate amplitude to the value R but substantial random drift in phase may occur.

Superimposed on the random walk of the phasor end, there will be a circular drift proportional to the departure of f from the exact center frequency of the source spectrum.

On the other hand, if the radiation studied is coherent, the phasor end will be observed to execute a random walk statistically describable by a probability distribution

which, in most instances, will be a circular "mole-run" centered at the origin, but with vanishingly small probability at that origin [Fig. 1(b)]. Any radial cross section of this mole-run will, in general, be approximately Gaussian. Radial excursions of the phasor end away from the center circle of the mole-run will resemble the excursions about zero of the thermal voltage on a condenser shunted by a resistance, for the regenerative property of the oscillator will tend to maintain constant the average length of the phasor. Conversely, circumferential excursions of the phasor end will assume the character of a random walk, the rate of which has been calculated in a former article.[2] A slow average rotation will be superimposed on this circumferential random walk, the rate of this rotation being proportional to the departure of f from the exact center frequency of the source spectrum.

It is essential to note that there is no fixed phase reference to which the phasor may be brought back with a restoring force proportional to the departure of the phasor from this phase reference. Thus, a probability distribution of the phasor end describable by a mole-hill at a distance from the origin (Fig. 2) is fundamentally impossible.

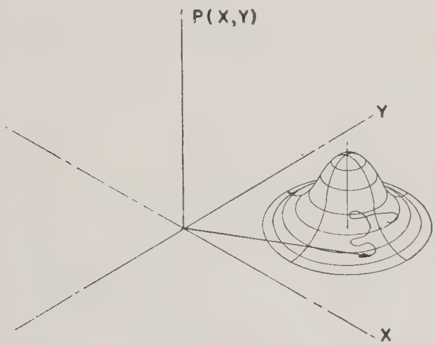


Fig. 2—This probability distribution is not a possible result in the limit as observation time approaches infinity since it would imply "phase-locking" with the hypothetical reference signal.

Experimental verification of the phasor's behavior for laser's outputs has not been obtained, but it may be speculated that when the radiation from two relatively stable lasers is caused to interfere on two photocells, with a quarter-wave difference between the two path differences from the two lasers to the two photocells, the photocells' outputs will represent a phasor, the phase of which is the phase difference of the two lasers. Whether or not the character of the probability distribution of the phasor end can be detected in such an experiment will depend upon whether the bandpass of the photocells' outputs exceeds or does not exceed the instantaneous frequency difference of the two lasers.

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[2] M. J. E. Golay, "Monochromaticity and noise in a regenerative electrical oscillator," Proc. IRE, vol. 48, pp. 1473–1477; August, 1960.

Note on the Probable Character of Intelligent Radio Signals from Other Planetary Systems*

Experiments are now being planned[1] to detect radio messages aimed at the earth from planets of other solar systems—experiments which few believe will succeed, but which few would not want to see tried.

The spectral location, as well as the character of these signals, constitute subjects for stimulating speculation.

It may well be surmised that these hypothetic signals will have a spectral location related in a simple manner to the 1421-Mc line of hydrogen, and since this line is located in a relatively noisy region of the cosmic radio spectrum, a location at either half or at twice its frequency appears fairly probable.

It may be surmised further that these signals will contain a coherent component[2] at as nearly half or twice the hydrogen frequency as available accuracy permits. Following this line of reasoning, we may expect the senders to correct this frequency for the component of motion of their transmitter in the direction of transmission with respect to the center of gravity of their planetary system, as well as for the gravitational potential of that transmitter with respect to their "local" outer space; corrections which we should also make for our receiver.

Even when so corrected, the coherent signal postulated here will have to be further corrected for the nebular drift between the two solar systems, unless of course it is already corrected at the transmitting end. Thus, it would appear that there may be four likely spectral regions within which a search should be made for coherent signals, before any intelligent messages can be received: at half or at twice the hydrogen frequency, and with or without correction for nebular drift.

The extent of each search region will be determined mostly by our—or their—uncertainty in nebular drift, and to a lesser extent by the relative error of sender and receiver in determining the exact hydrogen frequency or the various corrections indicated above. It must be noted that the two search regions at half the hydrogen frequency are only a quarter as wide as the two search regions at twice that frequency and that this factor of four is exactly compensated by the four-fold smaller directivity of a reflector of a given size at the lower frequency. When such factors as easier power generation and greater space coverage are considered, the two half-frequency regions appear more favorable.

The bandwidth of search, Δf, within each region is a matter for much conjecture, but once this bandwidth has been decided upon, the manner of search appears straight forward. The signals received are heterodyned with $\cos 2\pi$ ft and $\sin 2\pi$ ft, respectively, where f designates a frequency within the search region, and the two outputs are filtered by two low-pass RC filters with a time constant of the order of

$$RC = \frac{1}{\pi \Delta f}.$$

The filter outputs constitute the two components of a phasor, the behavior of the end of which should be studied for a time equal to a few times the time constant of the dual low-pass filters. If a "mole-run" tendency is detected for the statistical distribution of the phasor end, and if this tendency is confirmed by extending the time of search, the presence of intelligent transmission within the $f \pm \Delta f/2$ region will be ascertained with a probability of error which decreases exponentially with time.

Since, for a given transmitting power, the bandwidth of search Δf decreases with the inverse square of the distance, and since the time required for searching a given region increases with the inverse square of Δf, the fourth power law relating total time of search and distance will require that several Δf-wide regions be searched simultaneously, in order to cover the total search regions in a reasonable time.

It is of interest to note the basic difference between the search procedure outlined above, and the search procedure which consists in recording a narrow spectral region, forming the auto-correlation function of the recorded signals, and taking the Fourier transform in cosines of this auto-correlation function. This latter method serves to reveal the presence of extra spectral energy within narrow bands, but does not preserve phase information. Thus, a coherent signal which is slightly phase modulated becomes indistinguishable from a spectral line with a width equal to the frequency excursion, whereas the mole-run character of the phasor-end statistical distribution which can be detected with the method discussed above serves to establish with increasing certainty that coherent signals are indeed coherent.

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* Received by the IRE, February 8, 1961.

[1] O. Struve, "Astronomers in turmoil," *Phys. Today*, vol. 13, pp. 18–23; September, 1960.

[2] See preceding note in this issue.

VHF Satellite Signals Received at Extra Optical Distances*

Smyth Research Associates have been conducting an experimental program on low angle refraction of radio waves penetrating the atmosphere. Satellite transmitters at 108, 162, and 216 Mc have been used as signal sources in these studies. During the course of the measurements (June, July and August, 1960), it was found that on the majority of passes, signals were received while the satellite was well beyond the radio horizon. Out of a total of 49 cases when such "precursor" signals were sought, only 14 failed to yield signals for at least one minute beyond the radio horizon.

* Received by the IRE, March 3, 1961. This work was sponsored by Rome Air Dev. Ctr., ARDC.

In two cases, signals were first received when the satellite was 1700–1900 miles beyond the radio horizon point at orbit height. On these passes the satellite was roughly twice as far away as the distance to the horizon. The signal strength on these occasions was 11 db below the free space level on 216 Mc and 18 db below the free space level on 108 Mc. As can be seen in Fig. 1, there appears to be no distinct dependence of received signal strength on frequency. Even simultaneous data on 162 and 216 Mc show no consistent frequency dependence.

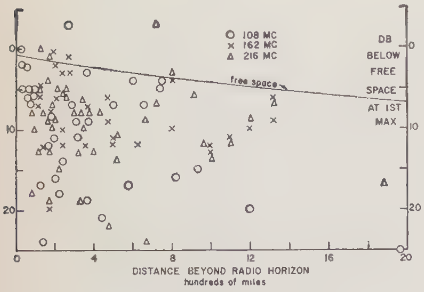

Fig. 1—Precursor signal levels.

The combination of low attenuation and no frequency dependence suggests a tropospheric ducting mechanism. Correlation of local radiosonde data with signal strength and maximum path length supports this view. The duct thickness and intensity in all cases when signals were observed, exceeded the trapping requirements for frequencies above 100 Mc.

Fig. 2 shows the data record for one of these cases, together with a scale of the great circle distance of the satellite below the radio horizon. The record was taken at the SRA Point Buchon field site on the coastline near San Luis Obispo, California, with a sea-interferometer system, giving rise to the pattern of maxima and minima above the horizon. Relatively minor tumbling nulls can also be seen superimposed on the interferometer trace.

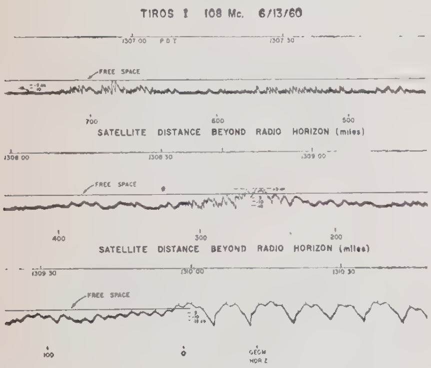

Fig. 2—Satellite precursor signals.

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WWV and WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible with respect to an improved United States Frequency Standard (USFS) since December 1, 1957.

The nominal broadcast frequencies should, for the purpose of highly accurate scientific measurements, or of establishing high uniformity among frequencies, or for removing unavoidable variations in the broadcast frequencies, be corrected to the value of the USFS, as indicated in the table. The corrections reported have been improved by a factor of three over those previously reported, by means of improved measurement methods based on LF and VLF transmissions.

The characteristics of the USFS, and its relation to time scales such as ET and UT2, have been described,[1] to which the reader is referred for a complete discussion.

The WWV and WWVH time signals are also kept in agreement with each other. Also they are locked to the nominal frequency of the transmissions and consequently may depart continuously from UT2. Corrections are determined and published by the U. S. Naval Observatory. The broadcast signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary. This new system was commenced on January 1, 1960. A retardation time adjustment of 20 msec was made on December 16, 1959; another retardation adjustment of 5 msec was made at 0000 UT on January 1, 1961.

WWV FREQUENCY
WITH RESPECT TO U. S. FREQUENCY STANDARD

1961 February	Parts in 10^{10}†
1	−150.7
2	−150.4
3	−150.5
4	−150.8
5	−150.9
6	−151.0
7	−151.1
8	−150.5
9	−149.8
10	−149.6
11‡	−148.0
12	−149.3
13	−149.6
14	−150.0
15	−150.1
16	−150.4
17	−150.7
18	−150.9
19	−150.8
20	−150.6
21	−150.5
22	−150.2
23	−150.2
24	−150.3
25	−150.4
26	−150.2
27	−150.3
28	−150.1

† A minus sign indicates that the broadcast frequency was low. The uncertainty associated with these values is $\pm 5 \times 10^{-11}$.

‡ The frequency was decreased 1×10^{-10} on February 11, 1961.

NATIONAL BUREAU OF STANDARDS
Boulder, Colo.

* Received by the IRE, March 17, 1961.

[1] "National Standards of Time and Frequency in the United States," PROC. IRE, vol. 48, pp. 105–106; January, 1960.

A Ruby Laser with an Elliptic Configuration*

A ruby laser with a new configuration was successfully operated at 1/15 input threshold energy of previous lasers.[1,2] This decrease in pump energy was accomplished by a new configuration which efficiently focuses the pump radiation into the ruby rod.

The laser is an optical maser[3] and is the only known device which can produce highly coherent radiation in the optical region. Fig. 1 shows details of the new device. Its most important elements are the specially processed ruby, the high-intensity discharge lamp which produces the pump power, and the cylindrical ellipse reflector which focuses the pump power into the ruby rod

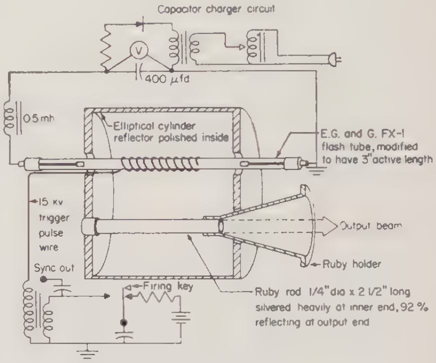

Fig. 1—Schematic of optical maser with elliptical reflector.

The ruby is an Al_2O_3 crystal with chromium atoms replacing some of the aluminum atoms. A ruby which was 0.05 per cent Cr_2O_3 by weight was used. The end planes of the ruby rod are flat to a fraction of a micron and parallel to within 50×10^{-6} inch.

The ruby laser has a high-energy pump power requirement. For the efficient use of the pump radiation a focusing system must concentrate the pump energy into the ruby, and must do it without the radiation passing through the lamp since an emission lamp is an absorbing medium. This can be achieved with an elliptical cylinder configuration with the lamp along one focal line and the ruby along the other. In order to have sufficient gain per reflection, a *long* cylindrical ruby is used to offset losses inherent in the reflecting coatings. Therefore, a *cylindrical* ellipse is a good configuration. The new laser was operated at 150 joules threshold pump energy as opposed to the 2300 joules previously required.

This configuration has other advantages. Cooling of either the ruby or the lamp or both can be achieved in a variety of ways,

* Received by the IRE, March 16, 1961.

[1] T. H. Maiman, "Stimulated optical radiation in ruby," *Nature*, vol. 187, pp. 493–496; August, 1960.

[2] R. J. Collins, D. F. Nelson, A. L. Schawlow, W. Bond, C. G. B. Garrett, and W. Kaiser, "Coherence, narrowing, directionality, and relaxation oscillations in the light emission from ruby," *Phys. Rev. Letters*, vol. 5, pp. 303–305; October, 1960.

[3] J. R. Singer, "Masers," John Wiley and Sons, Inc., New York, N. Y.; 1959.

such as circulating a coolant or using filters on the reflector walls or along the minor axis of the ellipse to protect the ruby from unwanted pump radiation. A cryostat can be used outside the elliptic cylinder to cool the ruby. Thus, continuous (CW) operation of this laser configuration may become feasible if sufficiently intense light sources can be found. Other details are given in Fig. 1.

Fig. 2 shows the energy levels and associated parameters of the chromium atoms in Al_2O_3 pertinent to the laser. These atoms are excited to the $^4F_2$ level by the radiation from the discharge lamp. For the production of coherent radiation ("laser-action"), it is

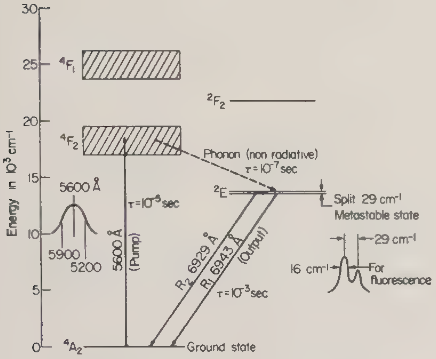

Fig. 2—Energy-level diagram of ruby.

necessary to highly overpopulate the ^2E metastable state. This is feasible primarily because of the broad energy width of the $^4F_2$ level and the large-lifetime τ_{21} compared to τ_{32}. The $^4F_2$ level is only one of several levels involved from which the atoms may decay to the metastable state.

Fig. 3 shows synchroscope pictures of the laser output vs time as a function of input

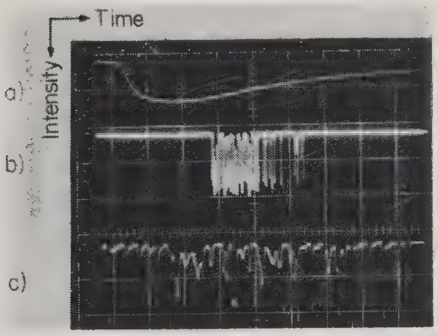

Fig. 3—(a) Fluorescence when input is 144 joules, gain 500, sweep 100 μsec/cm. (b) Laser output when input is 222 joules, gain 1, sweep 100 μsec/cm. (c) Laser output when input is 222 joules, gain 1, sweep 10 μsec/cm delayed 360 μsec from start of flash.

energy. At threshold input energy for laser action, the output radiation forms a parallel beam which produces a highly concentrated bright spot on the screen. The output radiation below threshold input energy comes from spontaneous emission and spreads in all directions and is barely visible. Further experiments are in progress to investigate the mechanism of the pulses[4] shown in Fig. 3(c). The energy bandwidth of the laser output as measured with a Fabry-Perot interferometer was found to be in the vicinity of 0.06 cm^{-1}. The details of these measurements will be reported later.

A self-contained portable version of this new laser has been constructed and weighs only 19 pounds, including the transistorized battery-operated power supply.

M. Ciftan
Microwave and Power Tube Division
C. F. Luck
C. G. Shafer
H. Statz
Research Division
Raytheon Company
Waltham, Mass.

[4] H. Statz and G. deMars, "Transients and oscillation pulses in Masers," in "Quantum Electronics Symposium," C. H. Townes, Ed., Columbia University Press, New York, N. Y.; 1940.

C-Band Nondegenerate Parametric Amplifier with 500-Mc Bandwidth*

A *C*-band nondegenerate parametric amplifier has been developed that has the following characteristics:

Signal frequency:	5.3 kMc
Idler frequency:	9 kMc
Pump frequency:	14.3 kMc
Instantaneous bandwidth:	500 Mc
Gain:	10 db
Noise figure:	3 db (including circulator loss).

A schematic diagram of the amplifier, which uses two commercially available silicon pill varactor diodes in a balanced circuit configuration, is shown in Fig. 1.

To obtain a large gain-bandwidth product simultaneously with a low noise figure and good operational stability, a circulator is used to interconnect the antenna, amplifier and post-receiver. Transformer T_1 changes the characteristic impedance of the circulator to the lower value required to obtain the necessary gain and bandwidth. The two varactor diodes are effectively in parallel in the signal circuit, and inductance L_s resonates with their net capacitance at the signal frequency. The parallel *L-C* combination in the signal circuit is used only when a double-tuned response is desired.

The equivalent idler circuit consists essentially of the two varactor diodes in series, since very little idler power flows in the signal circuit (because of the balanced circuit configuration) or in the pump circuit (because the pump power is applied through a high-pass filter). The zero-bias capacitances of the two diodes are matched within about 10 per cent, and each diode with its associated parasitic reactance is very nearly self-resonant at the idler frequency of 9 kMc. Thus, the idler circuit bandwidth is about the widest obtainable without additional resistive loading (which would degrade the amplifier noise performance).

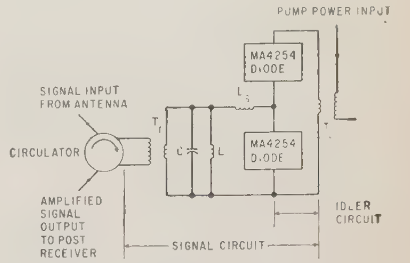

Fig. 1—Schematic diagram of *C*-band nondegenerate parametric amplifier.

Fig. 2 shows the measured gain-frequency characteristic of the amplifier under single-tuned conditions. As shown, a bandwidth of 500 Mc and a peak gain of 10 db are obtained. The noise figure of the amplifier-circulator combination is 3 db at the midband frequency of 5.3 kMc. The noise figure varies over the pass-band by about ± 0.5 db, being lower below midband frequency and higher above midband frequency mainly because of the changing ratio of signal-to-idler frequencies.

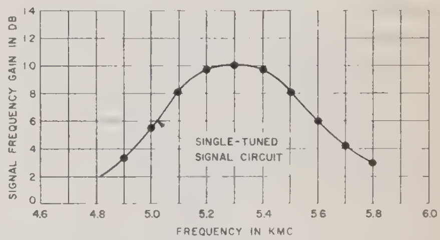

Fig. 2—Measured gain characteristic of *C*-band nondegenerate parametric amplifier.

The diodes are operated at zero bias and require a total pump power of about 130 mw. The amplifier gain varies by only ± 0.5 db for a ± 1-db change in pump power and is very insensitive to antenna, post receiver, or circulator impedance variations. Two of the single-tuned amplifier-circulator combinations can readily be cascaded to provide 20-db gain with an over-all bandwidth of 350 Mc; even wider bandwidths are possible with the same peak gain if the tuning of the amplifiers is slightly staggered. Further theoretical and experimental work is presently in progress to obtain much wider amplification bandwidths through the use of double-tuned signal and idler circuits.

The author gratefully acknowledges helpful discussions with J. C. Greene and J. J. Whelehan of AIL during the course of this work.

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Backward-Wave Radiation from Periodic Structures and Application to the Design of Frequency-Independent Antennas*

One characteristic common to a large number of successful frequency-independent log-periodic and log-spiral antennas is that the radiation is directed toward the apex or feed point of the structure.[1–3] In attempting to explain this property, and more generally to understand the operation of log-periodic antennas, it was found useful to think of the antenna as a locally periodic structure whose period varies slowly, increasing linearly with the distance to the apex.

It is well known that a number of diffracted beams are produced when a plane wave is diffracted by an infinite plane grating. Each beam corresponds to a space harmonic which has a phase constant $\beta_n=\beta_0-(2n\pi/a)$ along the grating where a is the period of the grating. The wave characterized by β_0 is the fundamental wave. The same description applies to a wave guided along the grating and having a phase constant β_0 in that direction. For example, Fig. 1 shows a dielectric slab bounded on one side by a conductor and on the other side by a grating. If this supports a slow wave ($\beta_0>k=\omega\sqrt{\mu\epsilon}$) and if the period a is small, $p=2\pi/a$ will be large and all the $\beta_n=\beta_0-np$ (for n positive or negative integer) will be larger than k. None of the diffracted beams has a real direction of propagation which means that no radiation occurs for the infinite structure. If now the spacing is increased, p will become smaller, the point β_1 (and all the points β_n) will move toward β_0 and eventually β_1 will become less than k. When this happens, the diffraction occurs in the negative direction (backward wave). If

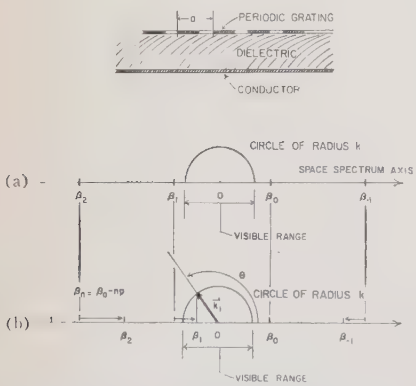

Fig. 1—(a) When a is small, $p=2\pi/a$ is large, no radiation occurs. (b) When a increases, the points β_n move as indicated. In the case shown here, one wave is in the visible range.

the spacing is further increased the points β_2, then β_3, and so on will enter the circle of radius k and several beams can result, radiated at angles from the z direction which may be obtained from the construction shown in Fig. 1.

$$\cos\theta_n=\frac{\beta_n}{k}\quad(|\beta_n|<k).$$

The increase in spacing may occur progressively along a single structure as in a log-periodic grating. As the fundamental wave progresses, it will successively reach points where the various conditions outlined above will occur. If the rate of increase is slow, a section of the structure about these points will behave as if it were periodic with the local period. As the fundamental wave reaches a radiation region, it will naturally be attenuated (because of this radiation) and the local value of β_0 may also be somewhat modified by this loading of the transmission medium. If the coupling is strong enough and if the detail design of the structure favors the backward wave, the radiation will cause attenuation of the fundamental wave such as to remove all the incident energy before the conditions for this radiation change too radically. The pattern will be approximately that of the periodic structure having the local value of the period. When the frequency is increased, the region where backward-wave radiation occurs (active region) will move but the pattern will be substantially the same.

These considerations have been applied and tested on periodic zig-zag and helical wires. The phase delay of the fundamental wave is approximately given by the phase constant, $\beta_0=k\csc\gamma$, $\gamma=$pitch angle. A zig-zag antenna was constructed which, according to the theory, should radiate in the backfire direction at 1500 Mc. Below this frequency, additional gain would be expected due to excess phase shift between cells. Above this frequency, the maximum radiation should scan away from backfire. The measured patterns are shown in Fig. 2. The same tests were made on a bifilar helix and similar results were obtained. The zig-zag and helical antennas operating in the backfire mode have smaller cross sections in terms of the wavelength than the conventional endfire zig-zag and helical antennas.

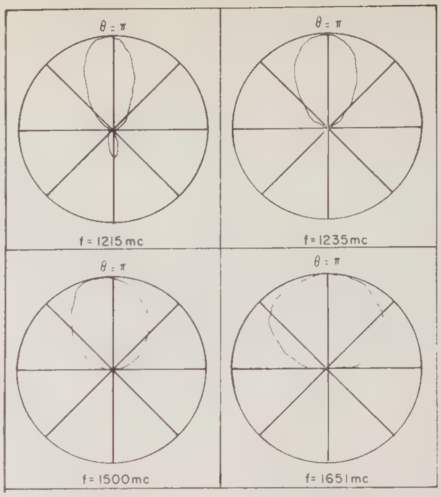

Fig. 2—E-plane radiation patterns of a backward-wave bifilar zig-zag antenna.

A number of new backward-wave structures are now proposed following the foregoing basic considerations. For example, a periodic monopole array over ground may be fed from a helical delay line. Typical radiation patterns of this type of periodic structure are shown in Fig. 3. The correspondence between the measured patterns and the predictions of the simple theory is good. The achievement of backward-wave radiation from an array of slots in a ground plane depends upon obtaining the proper slow phase velocity in a waveguide or line used to excite the slots. The coaxial guide with a helical center conductor seems most likely to operate satisfactorily.

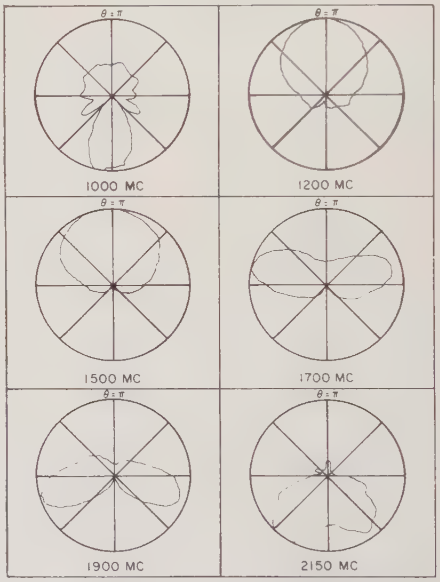

Fig. 3—H-plane radiation patterns of a periodic monopole array.

The analysis and design of frequency-independent antennas of both the log-periodic and log-spiral geometries is now underway using the foregoing principles. The design procedure is to first study, either theoretically or experimentally, a periodic structure made of the same elements. The frequency-independent design is then obtained from the periodic antenna by applying a linear taper, producing a log-periodic structure. The backward-wave bifilar helix is the periodic counterpart of the conical log-spiral antenna; the backward-wave zig-zag of the log-periodic zig-zag. Similarly, backward-wave periodic dipole, monopole and slot arrays may be converted into frequency-independent antennas by a linear taper.

The importance of these observations lies not only in the understanding that they contribute to the operation of frequency-independent antennas, but also in the new structures that they suggest. The basic ingredients for a frequency-independent antenna are a slow-wave transmission medium and a series of radiating elements satisfying the similarity condition and coupled to the transmission medium at points spaced in geometric progression. Proper design can produce a wave substantially radiated toward the feed point.

Backward-wave periodic structures without tapering are antennas which deserve at-

* Received by the IRE, February 3, 1961; revised manuscript received, February 16, 1961. The work described herein was supported in part by the Wright Air Dev. Div., ARDC, USAF, through Contract No. AF33 (616)-6079.

[1] D. E. Isbell, "Non-Planar Logarithmically Periodic Antenna Structures," Antenna Lab., Univ. of Illinois, Urbana, Tech. Rept. No. 30, Contract AF33 (616)-3220; February, 1958.

[2] R. H. DuHamel and F. R. Ore, "Logarithmically periodic antenna designs," 1958 IRE NATIONAL CONVENTION RECORD, pt. 1, pp. 139–151.

[3] J. D. Dyson, "The unidirectional equiangular spiral antenna," IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-7, pp. 329–334; October, 1959. Also, Antenna Lab., Univ. of Illinois, Urbana, Tech. Rept. No. 33, Contract No. AF33 (616)-3220; July, 1958.

tention for their own sake. The models which have been studied experimentally show very low sidelobes, high front-to-back ratio, and good bandwidth. Furthermore, the production of radiating waves from a periodic structure which is directly excited from a modulated electron beam appears to be a promising approach to the generation of microwave radiation.

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K_a-Band Ferrite Amplifier*

A ferrite parametric amplifier has been built and operated at a signal frequency of 20.5 kMc. It is believed that this is the highest frequency of operation of any ferrite parametric amplifier. The amplifier was operated with a pump power as low as 4 watts peak.

The amplifier used longitudinal pumping as suggested by Denton.[1] It consists of a signal cavity tunable by means of a sliding short coupled to a pump cavity also provided with a sliding short. Iris coupling was provided between the two cavities as well as between each cavity and its respective input guide. The two cavities are so arranged that the transverse field in the pump cavity (for the dominant mode) is perpendicular to the H plane in the signal cavity and thus to both transverse and longitudinal fields. The ferrite sample used was a highly polished single-crystal YIG sphere, 0.149 inch in diameter placed in the coupling region between the two cavities. A dc magnetic field was applied parallel to the transverse pump field and adjusted to resonance at the idle frequency. The rest of the microwave circuitry was conventional for one-port ferrite amplifiers and included an EH tuner in the pump line for impedance matching, a means of adjusting pump power level, band-pass filters in the signal line to block the pump power, and directional couplers for observation of reflected signal power and incident pump power. The pump power was obtained from a Microwave Associates MA210B magnetron operating at 34.8 kMc with a pulse duration of one microsecond. A 1000-cps square-wave amplitude modulated signal was used and when reflections were observed on a nonsynchronized oscilloscope two traces were displayed, one corresponding to periods with no signal input and the other corresponding to periods with a signal input. Since the pump, and therefore the amplifier, was pulsed, one picture of the oscilloscope face gave the amplifier response with and without signal and with and without pump. Measurements of net gain were obtained by comparing the output amplitude to a reference level obtained by replacing the amplifier with a short circuit. The amplifier output was attenuated sufficiently to match this reference level and the gain read directly from the attenuator.

The YIG sphere used was large enough so that many magnetostatic modes were capable of direct excitation. Because of propagation effects in the large sphere several distinct modes were observable in which there was negligible overlap. These are the modes which are most attractive for amplification. The idle mode was a magnetostatic mode near the bottom of the magnetostatic mode spectrum. The signal mode was a quasi-magnetostatic mode; that is, it was a ferrite mode with properties similar to magnetostatic modes but it lay outside the magnetostatic mode spectrum by about 4.5 kMc.

Experimental results with this amplifier are shown in Figs. 1 and 2. In Fig. 1, gain is plotted vs signal input power level, showing gain saturation for about one watt peak power output. In Fig. 2, small signal gain is plotted vs pump power. It is seen that with a pump power as low as 5 watts peak, a net gain of 10 db is obtained. It appears that gain drops to zero for a pump power of of 4 watts. However, the impedance matching used at the pump frequency was relatively inefficient, partly because of space limitations imposed by the large electromagnet used to furnish the dc field. It is anticipated that this threshold pump power can be reduced by several db with improvements in the matching structure.

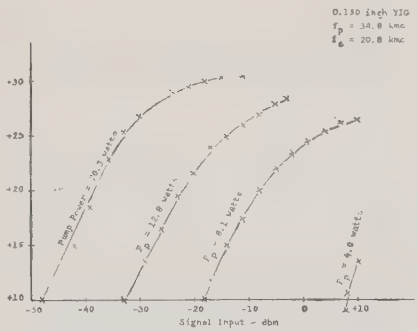

Fig. 1—K-band ferrite amplifier gain curves showing gain saturation.

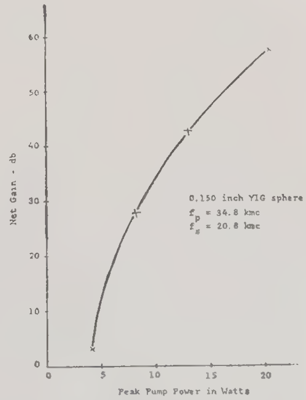

Fig. 2—Gain vs pump power for K-band ferrite amplifier.

No attempt was made to measure the noise figure of this amplifier since the magnetron pump source contributed considerable excess noise and the amplifier stability was relatively poor. It is anticipated that improvements in the pump circuit efficiency will allow the use of a klystron source, at least on a pulsed basis. Such improvements are presently being planned and noise figure measurements will then be made.

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[1] R. T. Denton, "A ferromagnetic amplifier using longitudinal pumping," PROC. IRE, vol. 48, pp. 937–938; May, 1960.

Equivalent Circuits for a Thermoelectric Converter*

When a thermoelectric converter is operated with constant heat input rather than at a constant temperature differential, new equivalent circuits are required to explain observed effects, particularly the transient response to step function disturbances. The dynamic behavior of a lightly loaded converter—such as might be used as a power supply for tunnel diodes—can be adequately described by a thermal statement that ignores Joulean heating, as well as the Thomson effect:

$$Q = (K + K_L)\Delta T + \Pi I + C_H \frac{d(\Delta T)}{dt} \cdots [\text{watts}]. \quad (1)$$

This is essentially a linear, nodal equation in which the time-dependent term is the heat flowing into or out of the heat capacitance C_H (watt-seconds per degree) as the result of any variation in the temperature, ΔT. Heat losses external to the thermoelements are represented by K_L added to the internal conductance K. The cold junctions are assumed to be held at a fixed temperature by an infinite heat sink, with respect to which the temperature ΔT of the hot junction is measured.

The physical model to which (1) applies is shown in Fig. 1, but if the Peltier heat ΠI is rewritten as $K_\pi \Delta T$ where $K_\pi = \Pi S(R + R_L)^{-1}$, the revised equation then applies directly to the circuit of Fig. 2. The steady-state response of the nodal temperature ΔT to a sinusoidal excitation $Q = Q_0 \cos \omega t$ is given by

$$\Delta T = \frac{Q_0}{K_\pi + K + K_L} R_e \left[\frac{\exp j\omega t}{1 + j\omega\tau}\right]$$

where

$$\tau = C_H(K_\pi + K + K_L)^{-1} \cdots [\text{sec}].$$

* Received by the IRE, January 9, 1961; revised manuscript received, January 19, 1961.

The corresponding output current may be written as

$$I = \frac{S\Delta T}{R + R_L}$$
$$= \frac{Q_0}{\Pi} \cdot \frac{R_\pi}{R_\pi + R + R_L} R_e \left[\frac{\exp j\omega t}{1 + j\omega\tau} \right]$$

where $R_\pi = \Pi S(K + K_L)^{-1}$, and Q/Π is a current source as shown in Fig. 3. The frequency dependence can be accounted for by the capacitance C also shown in Fig. 3, and it may be verified that the time constant is given by

$$\tau = \frac{R_\pi(R + R_L)C}{R_\pi + R + R_L} = \frac{\Pi SC}{K_\pi + K + K_L}.$$

Thus, the electrical equivalent of the thermal capacitance C_H, for a given device time constant, is $C = C_H(\Pi S)^{-1}$ [farads]. Since we may encounter resistance levels of hundredths of an ohm coexisting with time constants of hundreds of seconds, it will be commonplace to have values of 10^4 farads for the equivalent capacitance C.

Using the figure of merit $Z = S^2(RK)^{-1}$, and the Kelvin relation $\Pi = ST$, we can write the Peltier conductance K_π and the resistance R_π as, respectively,

$$K_\pi = \frac{R}{R + R_L} \cdot ZT \cdot K$$

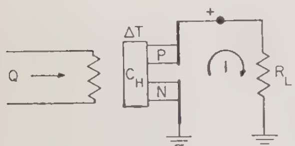

Fig. 1—Schematic of thermoelectric converter, emphasizing the thermal capacitance C_H at the hot junctions. The device also has the internal parameters, K the thermal conductance, R the electrical resistance, as well as the Seebeck and Peltier coefficients S and Π, respectively.

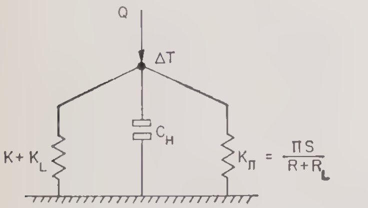

Fig. 2—Equivalent thermal circuit of converter, operating from a constant heat source Q. The equivalent conductance K_π represents Peltier heat pumping.

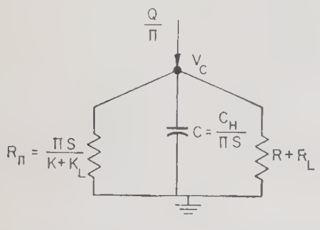

Fig. 3—Equivalent electrical circuit. The resistance R_π results from the constant heat input assumption, represented here by the current source Q/Π. Alternatively, the current source could be replaced by a voltage source $SQ(K + K_L)^{-1}$ in series with R_π.

and

$$R_\pi = \frac{K}{K + K_L} \cdot ZT \cdot R.$$

On the basis that the product ZT may have an upper boundary of the order of unity, K_π and R_π can, at most, be somewhat less than K and R, respectively.

A word of caution regarding efficiency: it must be defined as $\eta = I^2R_L/Q$, which reduces to

$$\frac{\Delta T}{T} \frac{R_\pi R_L}{(R_\pi + R + R_L)^2}$$

rather than

$$\frac{R_\pi R_L}{(R_\pi + R + R_L)(R + R_L)}$$

as one might deduce from Fig. 3. Despite low efficiencies, thermoelectric devices may provide a convenient means to simulate long time constants in control problems. The input could be supplied electrically either by Joulean heating or by Peltier pumping, giving rise to quadratic and linear input/output relations, respectively.

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A Method for Generating Signals of Arbitrary Yet Frequency-Independent Phase Differences*

When testing phase-measuring devices or synthesizing polyphase voltages, or in connection with auto-correlation procedures, function generation, etc., it is often useful to have a source of two or more signals at the same variable frequency, but with continuously adjustable, frequency-independent phase differences. It is the purpose of this note to describe one particularly simple method by which this can be achieved.

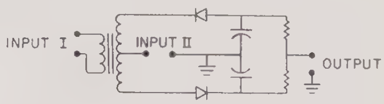

Fig. 1—Balanced-peak detector.

If, in the balanced-peak detector circuit shown in Fig. 1, a sine wave of large amplitude is applied to one of the input terminals and another sine wave of small amplitude and slightly different frequency is applied to the other, the output signal will be a sine wave at the difference frequency with an amplitude corresponding directly to that of the weaker input signal. If the two input signals are expressed as

$$E_1(t) = A_1 \cos(\omega_1 t + \Phi_1) \qquad (1)$$

* Received by the IRE, January 19, 1961.

and

$$E_2(t) = A_2 \cos(\omega_2 t + \Phi_2) \qquad (2)$$

where

$$A_1 \gg A_2, \qquad (3)$$

the output signal will be

$$e_1(t) = A_2 \cos[(\omega_1 - \omega_2)t + \Phi_1 - \Phi_2]. \qquad (4)$$

Here, ω_1 and ω_2 are the angular speeds of the two input signals, A_1 and A_2 are the respective peak amplitudes, and Φ_1 and Φ_2 correspond to some arbitrary phase angles. It is seen that by changing the phase of either one of the input signals, the phase of the output signal will change by the same amount. Consequently, in an arrangement such as is shown in Fig. 2, output signals of independently-adjustable phase differences can be obtained. By adjusting the frequency of the *undelayed* input signal, the frequency of these output signals can be varied over a wide range without changing their mutual phase relationships.

It should be noticed that it is the stronger of the two input signals that is being phase shifted. This implies that the changes in amplitude commonly associated with varying the phase of a signal will not have any effect on the output.

The actual operation of the arrangement shown in Fig. 2 has been experimentally tested. Two commercially-available signal generators were used as signal sources, and a continuously-variable delay line constituted the phase shifter. The two balanced detectors with the associated low-pass filters were identically constructed so as to cancel the effect of any phase shifts occurring therein. The fixed frequency of the one signal source was arbitrarily chosen to be 100 kc, implying that each microsecond of delay gave a 36° phase shift.

The result of the test is illustrated in Fig. 3. As can be seen there, the phase dif-

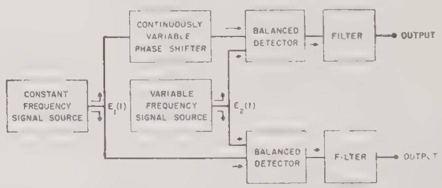

Fig. 2—Arrangement for obtaining signals of arbitrary, yet frequency-independent phase differences.

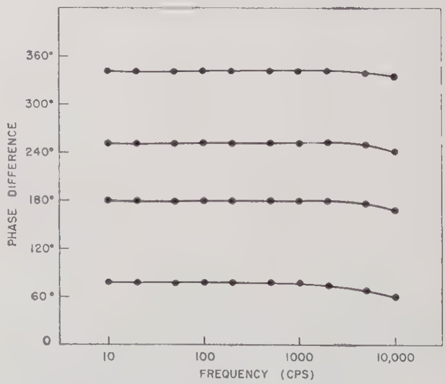

Fig. 3—Performance of the arrangement shown in Fig. 2.

ference between the two output signals at several different phase settings remained constant (within the $\pm 1°$ inaccuracy of the phase meter) over a range from 10 cps to 1 kc. Above 1 kc, a gradual change in phase was noticed—probably due to nonsymmetrical effects in the detectors and/or filters. No attempts were made to correct this condition. The limitations of the phase meter prevented any phase measurements below 10 cps. However, by observing Lissajous patterns on an oscilloscope, it appeared as if the phase relationship of the output signals remained constant down to essentially dc conditions. This, of course, was to be expected. The minimum operating frequency will be limited by the stability of the signal sources, rather than by the capabilities of the detectors and filters or of the method itself.

The author wishes to thank R. Geitka who constructed the apparatus and performed the testing involved.

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The Operation of Radio Altimeters Over Snow-Covered Ground or Ice*

The mode of reflection in the ground for the HF signals used in vertical incidence by ionospheric recording at the Royal Society Base at Halley Bay has been studied by Piggott and Barclay.[1] Several types of experiment show that the ice shelf is effectively transparent at HF and that most of the signal is reflected from the sea below the ice shelf rather than from the top of the shelf.

The purpose here is to draw attention to a possible source of danger when using radio altimeters in aircraft which may be particularly important when landing on flat surfaces covered by snow. Owing to a low density of even compacted snow, the signal from the air-snow interface is weak compared with that from the main reflecting interface. The actual circumstances, however, are not too common, and it is, therefore only necessary to give a very crude analysis which indicates the factors involved, and hence the occasions on which special precautions may be desirable.

When a plane wave passes through a dielectric material and is totally reflected at the lower surface of this material, the strength of the signal reflected from the upper surface increases with the dielectric constant of the material, and that from the lower surface decreases with it, the total energy being constant. The measurements at Halley Bay show that the partially reflected signal on frequencies near 3 Mc is 17 db weaker than the signal from the sea beneath the iceshelf.

Two factors determine the intensity of a signal reflected from a snow surface:

1) The mean density at the snow in the boundary layer in which reflection occurs, which is about a quarter of a wavelength thick.
2) The rate of change of density with depth, expressed in wavelengths in the medium, which determines whether the boundary should be considered to be sharp or diffuse. The partial reflection coefficient decreases as the boundary becomes more diffuse.

For altimeters operating near 100 Mc, wavelengths of about 3 meters, we may expect the partial reflection to occur within about 1 meter of the top surface.

The average density of snow at this depth is about 0.4 gm/cc so that the partial reflection loss at the air-snow surface is 23db. If the underlying reflector is the sea, the reflection at the top surface for a plane wave will be about 22 db weaker than that from the sea; if the lower reflection is from an iceshelf, the difference will be about 10 db. Frozen ground will lie between these limits. Thus the top surface reflection can be weak relative to that from the lower surface. This can be dangerous when the instrument is such that a weak reflection can be overlooked or is suppressed by a stronger signal reflected from similar ranges.

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This work forms part of the program of the British National Committee for the International Geophysical Year and the D.S.I.R. Radio Research Station, Slough, and is published by permission of the Director of Radio Research of the Dept. of Sci. and Industrial Res.

[1] W. R. Piggott and L. W. Barclay "The reflection of radio waves from an iceshelf," *J. Atmos. and Terres. Phys.*, in press; 1961.

High-Efficiency Variable-Reactance Frequency Multiplier*

The theory, design and performance of some maximum-efficiency variable-reactance frequency multipliers have been presented by the authors.[1,2] In view of the interest shown at the time of presentation, it is felt that some additional results would be welcomed. In this correspondence, the authors wish to report that a conversion efficiency of 40 per cent was obtained for a fifteen-times passive frequency multiplier using variable-capacitance diodes. Based on the disclosed theory, a quintupler was cascaded with a tripler. The results presented by the authors at the PGED meeting are reproduced here as shown in Fig. 1, where the efficiency vs harmonic number curve is compared with that reported by Leeson and Weinreb[3] and that of the ideal diode. Experimentally measured efficiencies confirmed theoretical estimation closely for harmonic order $N \leq 5$. A theoretical estimation for an eight-times multiplier is also indicated.

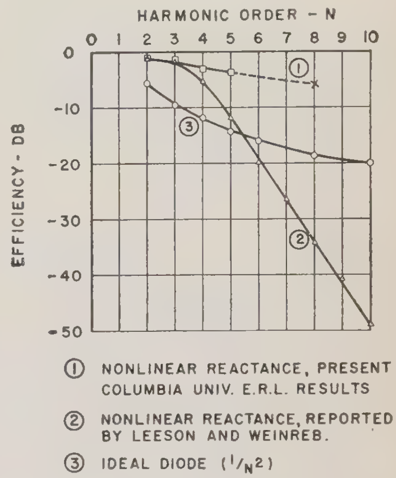

Fig. 1—Efficiency in db vs harmonic number.

In cascading the two multipliers, the following conditions have to be satisfied to assure high efficiency, where efficiency, η, is defined as

$$\eta = \frac{(\text{Output power at } Nf)}{(\text{Total input power}).}$$

1) The voltages across the nonlinear capacitors of both units have to be maximum without causing the diodes to conduct in the forward region.

2) The impedance levels of the two multipliers have to be matched to assure maximum power transfer from one unit to the next.

3) The tripler is used for the output stage because it is more effective in eliminating the undesired harmonics at the output.

The unit tested was operated with an input frequency of 0.84 Mc and an output frequency of 12.6 Mc. The input power to the device was approximately 10 mw, and the output power was 4 mw, which gave the efficiency of 40 per cent.

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* Received by the IRE, January 26, 1961. The research described in this paper stems from a project sponsored by the Air Res. and Dev. Command, USAF, and is administered by the Rome Air Dev. Ctr., Griffiss AFB, N. Y. under contract AF 30(602)-2233.

[1] T. Utsunomiya and S. Yuan, "Theory, Design and Performance of Maximum Efficiency Variable Reactance Frequency Multiplier," Electronics Res. Labs., School of Engrg., Columbia University, New York, N. Y., Tech. Rept. No. T-1/164, June 15, 1960.

[2] T. Utsunomiya and S. Yuan, "Theory, Design and Performance of Maximum Efficiency Variable Reactance Frequency Multiplier," Presented at the Annual Electron Devices Meeting of 1960, Washington, D. C.

[3] D. B. Leeson and S. Weinreb, "Frequency multiplication with nonlinear capacitors—a circuit analysis," Proc. IRE, vol. 47, pp. 2076–2084; December, 1959.

Parametric Up-Converter Tunable Over an 18:1 Frequency Band*

The operation of parametric amplifiers over frequency bands of an octave or more has not been extensively reported in the literature and indeed these devices have come to be regarded as essentially narrow-band amplifiers with typical frequency bandwidths of 10 per cent to 20 per cent.

Theoretical considerations did not indicate that these restricted bandwidths were necessarily associated with the upper sideband up-converter and an experimental investigation was carried out to explore the bandwidth limitations over wider frequency ranges in a device of this type. Signal frequency was varied over the range 100–1800 Mc and the pump frequency over the range 8500 to 6800 Mc to give a constant upper sideband frequency of 8600 Mc.

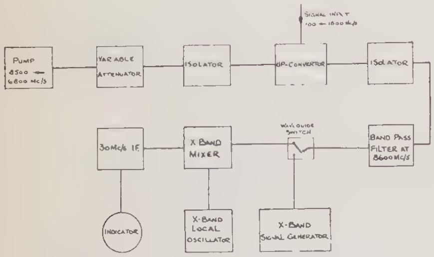

Fig. 1—Block diagram of experimental arrangement for up-converter measurements.

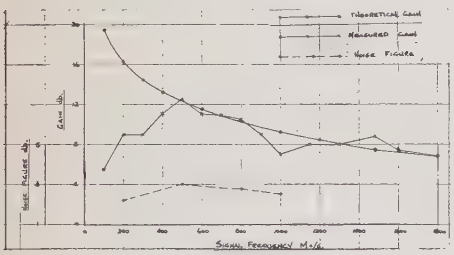

Fig. 2—Gain of up-converter.

With the amplifier illustrated in Fig. 1 gain has been obtained over the signal frequency range of 100–1800 Mc by varying only the pump and signal frequencies; no other tuning adjustments to the circuit were necessary. Particular care was taken to utilize an essentially aperiodic signal input circuit. The curves of measured gain and noise figure against frequency are shown in Fig. 2 together with the curve of the theoretical gain based on the Manley-Rowe frequency ratio relationship. It will be noted that the measured gain approaches the theoretical gain over the central and higher portion of the frequency range, while over the lower part of the frequency range the measured gain is lower than the theoretical curve. In the lower frequency region this is due largely to a poor match at the signal input. Thus, over the band 100–1800 Mc the band limits are set by the poor input match at the lower end and by fundamentally low gain due to the small frequency ratio at the upper end.

The accuracy of the gain measurements, dependent on the absolute calibration of two signal generators operating at the upper sideband and signal frequencies, is estimated to be ± 2 db. The noise figure measurements are included as an indication of the approximate noise level which is being investigated further. The pump power required was 250 mw. The diode used was a G.E.C. Type VX3333 with a cutoff frequency of 87 kMc.

Particular care was taken in the design of the amplifier to exclude the pump and the lower sideband signals from the output by means of a band-pass filter which passed only the upper sideband frequency. Ferrite isolators were used in the pump input and signal output waveguides to reduce the effects of unwanted reflections on the fields due to the pump and the two sidebands at the diode. This isolation was important in obtaining the wideband performance of the amplifier.

With the pump adjusted for amplifier gain at a signal frequency of 1000 Mc the rejection by the amplifier of input signals at frequencies which were submultiples of 1000 Mc is shown in Fig. 3. This characteristic is of importance when considering the level of spurious responses associated with the amplifier.

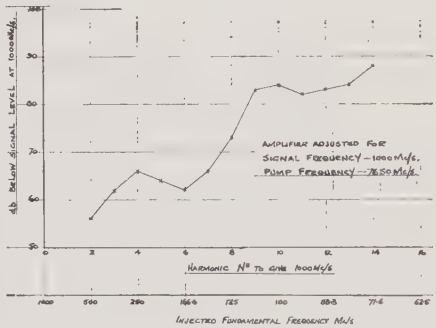

Fig. 3—Harmonic rejection of up-converter.

With the pump at a fixed frequency and the signal frequency varied, the gain was measured to determine the bandwidth over which it was obtained. This bandwidth was, as expected, wide and was found to be in excess of 150 Mc, measurement being restricted only by the tuning characteristics of the particular X-band superheterodyne receiver used as a detection device.

The above results do not necessarily represent the ultimate limits of operation of an upper sideband up-converter over the wide band of signal frequencies. They illustrate only the present state of the development of a wide-band amplifier and are considered to be of sufficient significance and general interest to warrant publication at this interim stage. Investigation of the amplifier characteristics continues.

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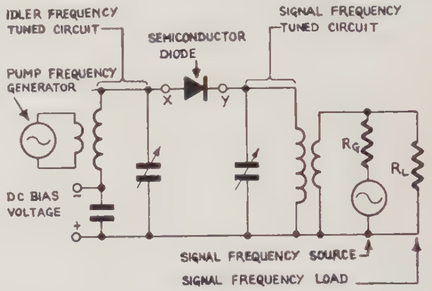
* Received by the IRE, January 10, 1961.

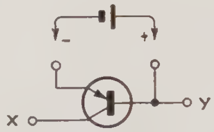
Parametric Amplification by Charge Storage*

When semiconductor diodes are used for parametric amplification, it is usually assumed that it is the variable depletion layer capacitance that provides the amplification. It is the purpose of this letter to point out that parametric amplification and subharmonic oscillation[1] can also be obtained from the effects of charge storage in diodes.[2]

A typical circuit for achieving parametric amplification at relatively low frequencies is shown in Fig. 1(a) and the theory of operation of this circuit in terms of the variable depletion layer capacitance has been extensively studied; however, as will be explained, this circuit can also be shown to provide parametric amplification by charge storage.



(a)



(b)

Fig. 1.

In studying the effects of charge storage on parametric amplification, transistors have been found useful, since the quantity of charge in the base region can be controlled, to some extent, by the emitter or collector. Now, if in the circuit of Fig. 1(a) we use the base collector junction of a transistor instead of a diode, as indicated by Fig. 1(b), then parametric amplification can be obtained due to the variable depletion layer capacitance. However, if the collector junction is driven into forward conduction during part of the pump cycle, and the pump frequency is not too low, then charge storage will occur; *i.e.*, the charge of minority carriers built up in the base region during forward conduction will allow the junction to continue conducting for part of the time during which the pump voltage reverse biases the diode. This charge storage can be almost completely eliminated by connecting a bat-

* Received by the IRE, January 30, 1961.

[1] Subharmonic oscillation due to charge storage has also been reported by W. D. Ryan, "Frequency division by carrier storage," *Electronic Engrg.*, vol. 33, pp. 40–41; January, 1961.

[2] The use of charge storage to provide parametric amplification and subharmonic oscillation is implied by D. Leenov and A. Uhlir, "Generation of harmonics and subharmonics at microwave frequencies with *p-n* junction diodes," Proc. IRE, vol. 47, pp. 1724–1729; October, 1959.

tery directly between emitter and base, so as to reverse bias the emitter junction. This has the effect of sweeping nearly all the minority carriers out of the base region within the transit time, so that, provided the pump frequency is not too high, there will be effectively no minority carriers left in the base to provide conduction during reverse bias. The connection of the battery has no effect on the depletion layer capacitance, however.

In all experiments carried out on the circuit of Fig. 1(a) using a transistor as described above, it has been found easy to obtain parametric amplification or subharmonic oscillation, provided the battery is disconnected and the base collector junction is driven into forward conduction. However, if the battery is connected, or the junction not driven into forward conduction, then the amplification or oscillation is greatly reduced, whence it is concluded that charge storage is principally responsible for the parametric effects observed.

In some of the parametric amplifiers that have been reported, the diodes are described as self biased[3] or zero biased,[4] in which cases the diode must be driven into forward conduction during part of the pump cycle; this must also occur if any rectified current is obtained. It thus seems probable that the amplification obtained in many parametric amplifiers depends, at least in part, on charge storage.

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[3] E. M. T. Jones and J. S. Honda, "A low-noise up-converter parametric amplifier," 1959 IRE WESCON CONVENTION RECORD, pt. 1, pp. 99–107.
[4] M. Uenohara, "Noise considerations of the variable capacitance parametric amplifier," PROC. IRE, vol. 48, pp. 169–179; February, 1960.

Approximate Method of Calculating Electromechanical Coupling Factor*

Usually, the electromechanical coupling factor k of the piezoelectric material is calculated from the series resonance frequency f_s and the parallel resonance frequency f_p of the fundamental mode in an appropriate transducer by the following equation (notations following IRE standards on piezoelectric crystals[1] are used unless otherwise mentioned):

$$k^2/(1-k^2) = (1/p)(f_p^2 - f_s^2)/f_s^2. \quad (1)$$

Or in the first approximation, by putting $\Delta f = f_p - f_s$,

$$k^2 = (2/p)(\Delta f/f_s). \quad (2)$$

Eq. (1), however, is an approximate formula and an error increases with k as shown in Table I. The writer found a better approximate formula (18). This is simpler than (1), and an error may be neglected even when k reaches 0.7.

TABLE I
CALCULATED VALUES OF k_{31}

Calculated from \ $\Delta f/f_s$	0.1	0.2	0.3	0.5	1.0
Eq. (2)	0.497	0.702	0.860	1.111	1.507
Eq. (19)	0.474	0.641	0.755	0.907	1.110
Eq. (1)	0.454	0.593	0.678	0.779	0.887
Eq. (18)	0.463	0.614	0.710	0.828	0.962
Exact Eq. (21)	0.464	0.616	0.714	0.838	1.000

When the electric equivalent circuit of the vibrator is considered,

$$C_1/C_0 = 1/r = (f_p^2 - f_s^2)/f_s^2, \quad (3)$$

where C_0 is the shunt capacitance and C_1 is the motional capacitance in the fundamental mode. Generally, $C_0 \approx C^{Cl}$, where C^{Cl} is the "clamped" capacitance, and the factor $\chi(\approx 1)$ is introduced[1] which takes into account the influence of the other modes of the vibrator excitable with the same electrode arrangement.

$$C^{Cl}/C_0 = \chi. \quad (4)$$

For the specimen with a single series of modes as fulfilled, *e.g.*, for the length-extensional mode of a bar, the contour-shear mode of a square plate, and the radial mode of a disk, the general electric equivalent circuit, which is correct in all the frequency range, is represented by connecting, in parallel, C^{Cl} (C_1 and L_1 in series), (C_2 and L_2 in series), $\cdots$ (C_∞ and L_∞ in series), the resistance terms being neglected. Namely, the admittance Y of the specimen is as follows:

$$Y = j\omega C^{Cl} + \sum_{n=1}^{\infty} \frac{j\omega C_n}{1-\omega^2 L_n C_n}. \quad (5)$$

That (5) is an exact solution is proved as follows: generally, Y is solved rigorously from the theory of vibrations as

$$Y = j\omega C^{Cl}\left\{1 + \frac{k^2}{1-k^2}\,\frac{1}{f(\omega)}\right\}, \quad (6)$$

and

$$f(\omega_{Rn}) = 0, \quad (n = 1, 2, 3, \cdots); \quad (7)$$

where ω_{Rn} is the resonance angular frequency of the nth overtone mode. By Mittag-Leffler's theorem,[2] $1/f(\omega)$ is developed as

$$\frac{1}{f(\omega)} = \sum_{n=1}^{\infty} \frac{p_n}{1-(\omega^2/\omega_{Rn}^2)}, \quad (8)$$

where p_n is a constant, and especially

$$p_1 = p. \quad (9)$$

Substituting (8) in (6), we obtain

$$C_n = p_n \frac{k^2}{1-k^2} C^{Cl}, \quad (10)$$

$$L_n = 1/(\omega_{Rn}^2 C_n). \quad (11)$$

Now, at the frequency near the resonance of the fundamental mode,

$$\omega^2 L_n C_n \ll 1 \quad (n = 2, 3, 4, \cdots), \quad (12)$$

and we can approximate as

$$C_0 = C^{Cl} + \sum_{n=2}^{\infty} C_n = C^F - C_1, \quad (13)$$

where C^F is the "free" capacitance, namely

$$C^F = C^{Cl} + \sum_{n=1}^{\infty} C_n = C^{Cl}/(1-k^2). \quad (14)$$

From (13), (14), (10) and (9), we obtain

$$C_0/C^{Cl} = (1-pk^2)/(1-k^2) \quad (15)$$

$$C_1/C^{Cl} = pk^2/(1-k^2). \quad (16)$$

Accordingly, by comparing (15) and (16) with (3), we obtain

$$k^2 = \frac{1}{p(1+r)}, \quad (17)$$

or

$$k^2 = (1/p)(f_p^2 - f_s^2)/f_p^2. \quad (18)$$

And when $(\Delta f/f_p)^2$ is neglected,

$$k^2 = (2/p)\Delta f/f_p. \quad (19)$$

Eq. (18) is a new approximate formula, and this approximation is, from (15) and (4), equivalent to

$$\chi = (1-k^2)/(1-pk^2), \quad (20)$$

while (1) corresponds to $\chi = 1$.

In Table I, the calculated values by (1), (2), (18) and (19) are compared with true values calculated by (21),[3] in the case of the length-extensional mode where $p = 8/\pi^2$.

$$\frac{\pi}{2}\left(\frac{f_p}{f_s}\right)\tan\left\{\frac{\pi}{2}\left(\frac{f_p}{f_s}-1\right)\right\} = \frac{k_{31}^2}{1-k_{31}^2}. \quad (21)$$

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* Received by the IRE, January 30, 1961.
[1] IRE Piezoelectric Committee, "IRE standards on piezoelectric crystals, 1958," PROC. IRE, vol. 46, pp. 764–778; April, 1958.
[2] E. T. Whittaker and G. N. Watson, "Modern Analysis," Cambridge University Press, Cambridge, Eng., pp. 134–136; 1935.
[3] W. P. Mason, "Piezoelectric Crystals and their Application to Ultrasonics," D. Van Nostrand Co., Inc., New York, N. Y., pp. 61–68; 1950.

A Simple Method of Generating Nanosecond Pulses at X Band*

Present methods of generating nanosecond RF pulses use high-speed switches to amplitude modulate microwave energy. One approach uses fast semiconductor diodes,[1,2] while Beck's[3] utilizes the TWT as a switch. A TWT can also be used in a nonlinear feedback loop to form a regenerative pulse generator.[4] The technique to be described, based on the impulse response of a TWT, is a simple means of generating short RF pulses.

Theoretically, the Fourier spectrum of dc pulses extends out to infinite frequency. For pulses of the order of 1 nsec in width and a few volts amplitude, the components in the X-band region are well above thermal noise at the input of a 50-Ω system. When such short pulses are fed directly into an X-band TWT serving as a band-pass amplifier, RF output pulses comparable in width to the dc input pulses are obtained.

The equipment is shown in Fig. 1; the TWT amplifier consisting of two tubes in cascade gave an estimated 0.5-watt peak-pulse power. The dc pulser produced pulses of 1-nsec base width, rise time of 0.25 nsec and a PRR of 720 pps. The sampling oscilloscope (rise time 0.6 nsec) displays this pulse as shown in Fig. 2(a) and the detected output of the TWT as shown in Fig. 2(b). RF power at the crystal detector (IN23B) was kept low, resulting in square-law operation and a displayed pulse proportional to power. Detected pulse-rise time is about 1 nsec and base pulse width, 2 nsec. Observations on a 2000-Mc traveling-wave oscilloscope gave a rise time of 0.6 nsec and a base pulse width of 1.5 nsec.

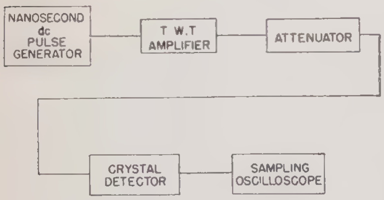
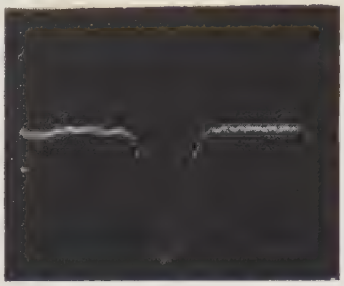


Fig. 1—RF pulse generator using a dc-pulse generator and TWT amplifier.

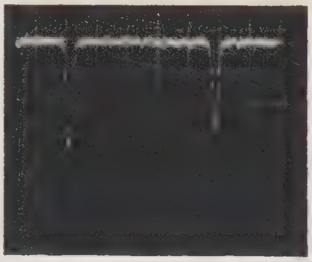
(a)



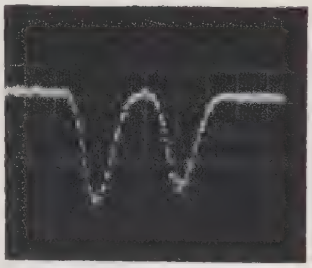
(b)

Fig. 2—(a) dc input pulse to TWT as observed on a 0.6 nsec rise time sampling oscilloscope. (b) Detected RF output pulse viewed on the same oscilloscope. Horizontal scale is 1 nsec per major division.

To study the use of the method in radar applications, the system of Fig. 1 was altered to include transmitting and receiving antennas. Echoes from various objects in the room were observed. Fig. 3(a) shows echoes from two plane metal surfaces 12 feet apart in range, the nearest being 15 feet from the two horn antennas. In Fig. 3(b) the surfaces are about 17 inches apart, the nearest being 40 feet from the antennas. Pulse amplitudes shown are not significant as the reflectors were positioned to give a display convenient for observation. Echoes from plane metal surfaces differing in range by six inches could be resolved easily.



(a)



(b)

Fig. 3—(a) Radar echoes from plane metal surfaces 12 feet apart in range and in line with the transmitting and receiving antennas (scale = 5 nsec/major divison). (b) Echoes from plane metal surfaces 17 inches apart (scale = 1 nsec/major division).

Because the traveling-wave tube is used as an amplifier, this technique is advantageous in setting up a sensitive short-pulse radar. A reflex scheme (Fig. 4) suggested by W. L. Haney uses the same TWT to amplify the received signal. Little additional equipment is required, and the system remains simple. With the setup of Fig. 4, the total gain of the system could not be used, as strong echoes from large objects in the laboratory caused oscillation. Even so, the sensitivity was improved about 25 db.

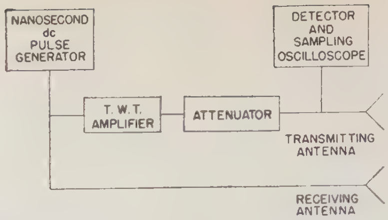


Fig. 4—Sensitive, short-pulse reflex radar.

An application of this method has been independently developed in this laboratory by S. G. Jones and T. H. Shepertycki for obtaining the impulse response of S-band radar receivers. Other applications contemplated for investigation include the location of troublesome reflections on antenna ranges and in "anechoic" rooms. Radar cross-section measurements of small objects and propagation delay studies are also possible.

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* Received by the IRE, January 23, 1961.

[1] K. J. S. Cave, W. Neu, and A. C. Sim, "A diode modulator for millimetre waves," *Proc. IEE*, vol. 106, pt. B, suppl. No. 16, pp. 759–761; May, 1959.

[2] C. A. Burrus, "Millimicrosecond pulses in the millimetre wave region," *Rev. Sci. Instr.*, vol. 28, pp. 1062–1065; December, 1957.

[3] A. C. Beck, "Waveguide investigation with millimicrosecond pulses," *Bell Sys. Tech. J.*, vol. 35, pp. 35–65; January, 1956.

[4] C. C. Cutler, "The regenerative pulse generator," Proc. IRE, vol. 43, pp. 140–148; February, 1955.

Whistlers Excited by Sound Waves*

Recently Lippmann[1] has mentioned the possibility that electromagnetic radiation in the low-frequency ranges as produced by a nuclear explosion could propagate as a "Whistler" along the magnetic field of the earth. Lippmann discusses the excitation of this whistler (see under No. 3). We are able to contribute to this discussion considerations which we have made under a more general aspect.[2]

The waves produced directly by the explosion must essentially be of longitudinal character, *i.e.*, they are sound waves. The whistler mode, however, corresponds to waves with essentially transversal polarization. Now Piddington[3] has shown that in the ionospheric plasma four types of electromagnetic waves with different characteristic polarizations can exist; two of them are essentially transversal and the other two essentially longitudinal in polarization. This is due to the influence of the magnetic field of the earth. Both transversal polarizations rotate in an opposite sense. In one sense of polarization, the electric vector rotates in the same way as do the positive ions in their free spiralling movement with the earth's magnetic field as axis; the other sense of polarization corresponds to the free spiral-

* Received by the IRE, January 25, 1961. This work was done under contract No. DA-91-591-EUC-1504 of the U. S. Army.

[1] B. A. Lippmann, "Bomb-excited whistler," Proc. IRE, vol. 48, pp. 1778–1779; October, 1960.

[2] K. Rawer and K. Suchy, "Longitudinal- und Transversal-Wellen im Lorentz-Plasma," *Ann. Physik*, vol. 3, pp. 155–170; March, 1959.

[3] J. H. Piddington, "The four possible waves in a magneto-ionic medium," *Phil. Mag.*, vol. 46, pp. 1037–1050; October, 1955.

ling movement of the electrons. In literature these polarizations are mostly designated as "ordinary" and "extraordinary" polarization; following Larenz[4] we prefer to call them "ionic" and "electronic," respectively. The two longitudinal polarizations have also a direct relation with the plasma components; they correspond to a kind of sound wave propagated in the ionic and electronic gas, respectively.

For wave frequencies which are considerably higher than the gyrofrequency of the ions, we can neglect the ionic sound wave. We have recently investigated the relations between the remaining three characteristic waves.[2] The following conclusions regarding the excitation of whistlers by sound waves in a plasma can be obtained:

In the inner ionosphere (below about 300 km), the plasma density increases nearly monotonically with height. This change causes a corresponding continuous change of the characteristic polarizations. The transversal waves which are of purely transversal polarization at the lower border of the ionosphere get a longitudinal component which increases more and more; as to the electronic sound wave which is purely longitudinal at the lower border of the ionosphere we find in the plasma a transversal component increasing more and more. Depending on wave frequency and magnetic field we may find a certain electron density for which the electronic sound wave and the electronic transversal wave have exactly the same polarization. These are the special conditions where an electronic sound wave can excite a transversal wave so that part of the sound wave energy coming from the explosion can be transformed into transversal electromagnetic wave energy.

Fig. 1 is reproduced from our quoted work;[2] it shows the square (w) of the refractive index as a function of the reduced plasma density $X=\omega_N^2/\omega^2$ (ω_N plasma pulsation, ω pulsation of the wave). The different curves are valid for the three different characteristic waves in the two limiting cases, *i.e.*, the transversal case ($\theta=\pi/2$) and the longitudinal case ($\theta=0$). In the latter case, the electronic sound wave reacts with the electronic transversal wave in the transition point which is found on Fig. 1 near $X=1$. We have shown in the quoted work that in the case of frequencies below the electronic gyrofrequency (for which case Fig. 1 is valid), the transition can only become important when the direction of propagation nearly equals that of the magnetic field. This latter condition is valid in the case of whistlers and our discussion shows that the transition will occur essentially in the neighborhood of that electron density for which the plasma frequency equals the wave frequency.

Another result concerning the absorption of the electronic sound wave can be obtained from our quoted work.[2] This absorption becomes stronger and stronger with increasing angle θ between the direction of wave propagation and that of the magnetic field. As compared with the absorption coefficient of the transversal waves, that of the electronic sound wave is greater by about a factor of $3m_ec_0^2/(5kT)=3.556\cdot10^9$°K$/T$. The absorption is very large so that only very strong sound waves are able to produce observable transversal electromagnetic waves. It seems quite reasonable that nuclear explosions at great altitude excite such strong sound waves.

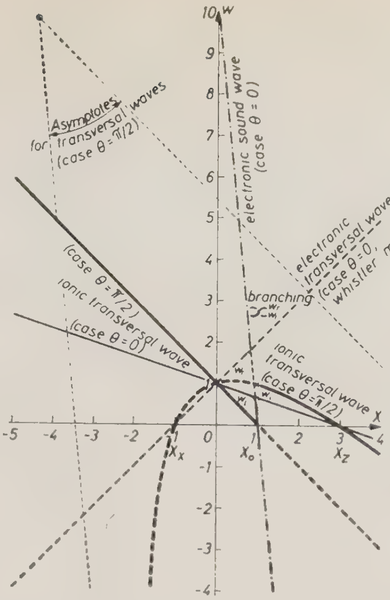

Fig. 1—Effective dielectric factor w ($\approx$square of the refractive index) as a function of the reduced electron density $X(=\omega_N^2/\omega^2)$ for wave frequency $\omega/2\pi$ = half of the electron gyrofrequency and $3m_ec_0^2/(5kT)=10$. Full lines $=w_l$ = ionic transversal wave, broken lines $=w_r$ = electronic transversal wave, point-dotted lines = electronic sound wave, heavy lines = transversal case ($\theta=\pi/2$), and thin lines = longitudinal case ($\theta=0$).

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Power Carried by the Cyclotron Waves and the Synchronous Waves on a Filamentary Electron Beam*

Waves on a filamentary electron beam in a transverse-field slow-wave circuit have been analyzed by A. E. Siegman,[1] who found the existence of the fast and slow cyclotron waves, as well as fast and slow synchronous waves, on such an electron beam. It is the purpose of this paper to show that the same result can be expressed in a different way by introducing the concept of the effective transverse current K and "the synchronous voltage," S, as well as "Chu's kinetic voltage," U. It is found that quite similar expressions for the beam admittance and the power of each wave component to those of the longitudinal electron beam can be obtained by this analysis.

The following equations are obtained from the force equation for electrons which are travelling to the z direction with velocity u_0 under dc conditions in an axial magnetic field B_0.

$$\frac{dU_-}{dt} - j\omega_c U_- = u_0 E_-$$
$$\frac{dU_+}{dt} + j\omega_c U_+ = u_0 E_+ \tag{1}$$
$$\frac{d}{dt}(U_- + jB_0 u_0 r_-) = u_0 E_-$$
$$\frac{d}{dt}(U_+ - jB_0 u_0 r_+) = u_0 E_+ \tag{2}$$

where the definitions of quantities are given as follows:

$U_\pm$ = Chu's kinetic voltage, $-(m/e)\, u_0 v_\pm$
$v_\pm$ = transverse velocity of the electrons in the beam
ω_c = cyclotron angular frequency
$r_\pm$ = transverse displacement of the electron beam
$E_\pm$ = transverse electric field applied to the electron beam
A_+ = positive circularly-polarized component of quantity A
A_- = negative circularly-polarized component of quantity A.

Now, we define a new parameter S which is named "synchronous voltage" because of the effective voltage for the synchronous waves (the author does not believe, however, that this name is the most suitable one).

$$S_- = U_- + jB_0 u_0 r_-$$
$$S_+ = U_+ - jB_0 u_0 r_+. \tag{3}$$

Eq. (2) can be rewritten by the definition in (3) as follows:

$$\frac{d}{dt} S_- = u_0 E_-$$
$$\frac{d}{dt} S_+ = u_0 E_+. \tag{4}$$

The effective transverse current K caused by the displacement of the electron beam from dc value is given as follows:[2]

$$K_\pm = j\omega\sigma_0 r_\pm, \tag{5}$$

where σ_0 is the total charge of the electron beam per transverse unit area. From (3) and (5) the following relationships are obtained:

$$K_- = -\frac{\omega}{\omega_c}\frac{|I_0|}{2V_0}(S_- - U_-)$$
$$K_+ = \frac{\omega}{\omega_c}\frac{|I_0|}{2V_0}(S_+ - U_+), \tag{6}$$

where V_0 is the dc accelerating voltage corresponding to the drift velocity u_0, and I_0 is the dc beam current ($|I_0| = -\sigma_0 u_0$).

[4] K. W. Larenz, "Beitrag zur Magneto-Hydrodynamik kompressibler Medien," presented at lecture in Bad Salzuflen, Germany; April, 1952 (unpublished).

* Received by the IRE, January 3, 1961.

[1] A. E. Siegman, "Waves on a filamentary electron beam in a transverse-field slow-wave circuit," *J. Appl. Phys.* vol. 31, pp. 17–26; January, 1960.

[2] H. A. Haus, "Electron Beam Waves in Microwave Tubes," in "Electronic Wave Guides," Polytechnic Press, New York, N. Y.; 1958.

The effective voltage and current for the cyclotron waves themselves can be obtained by putting $E_{\pm}=0$ and $S_{\pm}=0$ in (1) and (6) as follows:

$$U_- = U_{f-}e^{-j(\beta_e+\beta_c)z}\cdot e^{j\omega t}$$
$$U_+ = U_{s+}e^{-j(\beta_e-\beta_c)z}\cdot e^{j\omega t} \qquad (7)$$
$$K_- = \frac{\omega}{\omega_c}\frac{|I_0|}{2V_0}U_{f-}e^{-j(\beta_e+\beta_c)z}\cdot e^{j\omega t}$$
$$K_+ = -\frac{\omega}{\omega_c}\frac{|I_0|}{2V_0}U_{s+}e^{-j(\beta_e-\beta_c)z}\cdot e^{j\omega t}, \qquad (8)$$

where $\beta_e=\omega/u_0$, $\beta_c=\omega_c/u_0$. The power carried by the fast and slow cyclotron waves can be given by the generalized Chu's power theorem as follows:

$$P_f = \tfrac{1}{2}\,\mathrm{Re}\,[U_-K_-^*] = \frac{|I_0|}{4V_0}\frac{\omega}{\omega_c}|U_{f-}|^2$$
$$P_s = \tfrac{1}{2}\,\mathrm{Re}\,[U_+K_+^*] = -\frac{|I_0|}{4V_0}\frac{\omega}{\omega_c}|U_{s+}|^2. \quad (9)$$

The characteristic beam admittance for the small-signal quantities of the cyclotron waves is expressed by (7) and (8) as follows:

$$Y_0 = \frac{K_-}{U_-} = -\frac{K_+}{U_+} = \frac{\omega}{\omega_c}\frac{|I_0|}{2V_0}. \qquad (10)$$

This expression is very similar to that for the longitudinal beam (one-dimensional) given by the following equation.

$$Y_0 = \frac{\omega}{\omega_q}\frac{|I_0|}{2V_0}, \qquad (11)$$

where ω_q is the reduced plasma angular frequency.

The same equations as (7) and (8) for the synchronous waves can be obtained by putting $E_{\pm}=0$ and $U_{\pm}=0$ in (4) and (6):

$$S_- = S_{s-}e^{-j\beta_e z}\cdot e^{j\omega t}$$
$$S_+ = S_{f+}e^{-j\beta_e z}\cdot e^{j\omega t} \qquad (12)$$
$$K_- = -\frac{\omega}{\omega_c}\frac{|I_0|}{2V_0}S_{s-}e^{-j\beta_e z}\cdot e^{j\omega t}$$
$$K_+ = \frac{\omega}{\omega_c}\frac{|I_0|}{2V_0}S_{f+}e^{-j\beta_e z}\cdot e^{j\omega t}. \qquad (13)$$

Also, the power carried by the slow and fast synchronous waves are

$$P_s = \tfrac{1}{2}\,\mathrm{Re}\,[S_-K_-^*] = -\frac{|I_0|}{4V_0}\frac{\omega}{\omega_c}|S_{s-}|^2$$
$$P_f = \tfrac{1}{2}\,\mathrm{Re}\,[S_+K_+^*] = \frac{|I_0|}{4V_0}\frac{\omega}{\omega_c}|S_{f+}|^2. \qquad (14)$$

The characteristic beam admittance is given as (10).

It should be noticed that the power given by (9) and (14) is not the true transverse power (meaning, strictly speaking, the energy transfer due to the transverse fields), but the total power carried by each wave as pointed out by Siegman. This is because the effective transverse current K given by (5) is not the true transverse displacement current which should be

$$K_{\pm\mathrm{true}} = \left(j\omega + u_0\frac{\partial}{\partial z}\right)r_{\pm}\sigma_0 = v_{\pm}\sigma_0. \qquad (15)$$

And so,

$$K_{\pm\mathrm{true}} = \frac{|I_0|}{2V_0}U_{\pm} \text{ for the cyclotron waves}$$
$$K_{\pm\mathrm{true}} = 0 \text{ for the synchronous waves.} \qquad (16)$$

Therefore, the true power in the transverse direction, P_t, is

$$P_{t\,\mathrm{cyc}} = \tfrac{1}{2}\,\mathrm{Re}\,[U_{\pm}K_{\pm}^*{}_{\mathrm{true}}] = \frac{|I_0|}{4V_0}|U_{\pm}|^2.$$
$$P_{t\,\mathrm{syn}} = 0 \qquad (17)$$

P_ts for the fast and slow cyclotron waves are both positive and the same because of the kinetic power due to the cyclotron rotation of electrons in the transverse direction. P_ts for the synchronous waves are zero because there is no transverse motion of electrons for these waves.

The power in the longitudinal direction can be obtained by substracting P_t from the total power as follows:

$$P_{lf\,\mathrm{cyc}} = \left(\frac{\omega}{\omega_c}-1\right)\frac{|I_0|}{4V_0}|U_{f-}|^2$$
$$P_{ls\,\mathrm{cyc}} = -\left(\frac{\omega}{\omega_c}+1\right)\frac{|I_0|}{4V_0}|U_{s+}|^2$$
$$P_{lf\,\mathrm{syn}} = \frac{\omega}{\omega_c}\frac{|I_0|}{4V_0}|S_{f+}|^2$$
$$P_{ls\,\mathrm{syn}} = -\frac{\omega}{\omega_c}\frac{|I_0|}{4V_0}|S_{s-}|^2. \qquad (18)$$

In conclusion, the total power flow along the electron beam discussed here is given as

$$P = \tfrac{1}{2}Y_0[|U_{f-}|^2 - |U_{s+}|^2 + |S_{f+}|^2 - |S_{s-}|^2]. \quad (19)$$

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Parametric Variable-Capacitor Motor*

The problem of rotary motion with stimulance injected parametrically via a magnetic field has already been solved.[1] An active element appears in the electromagnetic network thanks to the total derivative of $d(Li)/dt$, but a transistor amplifier was included for stimulance enhancement. The underlying integro-differential equation with a Mathieu-Hill-type solution, covering the total electromagnetic system, is of the form

$$L\frac{di}{dt} + \left(R_P + \frac{dL}{dt}\right)i + S\int idt = e(t). \quad (1)$$

Two operators pertaining to spinning sinors can be identified, inviting the use of the Candy-Hollingworth operational equation solution with hyper-complex numbers, developed in England.[2] One operator is $d/dt \equiv j\omega = j2\pi f = j2\pi 60$ for the driving voltage, and the other is $d/dt \equiv j\Omega = j2\pi n_s$, where n_s is the rotational speed of the motor shaft in turns per second. R_P designates the positive or loss resistance of the total electromechanical system. It is now very logical to ask the following question: Would the dual of the network covered by (1) possibly yield an additional realizable physical system? If so, we can make a parametric electric motor based on the total differential of $d(Cv)/dt$, where $C=1/S$ is the variable capacitance. Such a motor would have much in common with the now popular semiconductor diode, variable capacitance, parametric microwave and wide-band amplifier. Replacing magnetic-field injected stimulance by electric-field injected stimulance, we obtain

$$C\frac{dv}{dt} + \left(G_P + \frac{dC}{dt}\right)v + \Gamma\int vdt = i(t), \quad (2)$$

where $\Gamma=1/L$ is the reciprocal inductance and G_P the positive or loss conductance of the total electromechanical system. With reference to both (1) and (2), transistor amplification may be added for enhanced stimulance, resulting in both cases in a practical brushless electric motor. Both (1) and (2) are linear equations, and rotation may therefore be predicted by the Nyquist stability theorem. It goes without saying that pendulum motion is equally well predicted, and that maintained pendulums may be built this way, provided the proper restoring force is introduced to change the rotary motion into oscillatory motion.[3]

Guided by theoretical speculations of the above nature, the writer carried out experiments with the Class-C amplifier shown in Fig. 1. While the explanation for the operation is given by the term dC/dt in (2), a more direct visualization of how the active element is introduced is obtained from the following reasoning. Let us assume that the moving vanes of the capacitor C_1 in Fig. 1 are on their way into the system of fixed vanes, and that the condition of resonance is approached. Accordingly, the alternating voltage across the capacitor C_1 experiences an increased amplitude, with an electric field attraction force appearing, which aids the movement of the vanes and thus provides a rotational force. All that is needed for practical motor operation is suitable programming of the source $i(t)$, so that self-commutation obtains. The necessary synchronization is indicated by the dotted arrow marked "SYNC" in Fig. 1. To test this idea, the writer removed the stops from the variable capacitor, reduced the friction in the bearings, and with an elementary form of synchronization, the rotor of the variable capacitor began to turn around; the variable capacitor was now an electric motor.[4] Several

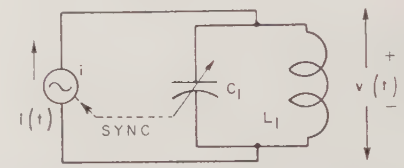

Fig. 1—Basic scheme of making the tuning capacitor in a tank circuit spin around so as to become an electric motor.

* Received by the IRE, January 30, 1961.

[1] H. E. Stockman, "Parametric oscillatory and rotary motion," PROC. IRE (Correspondence), vol. 48, pp. 1157–1158; June, 1960.

[2] C. J. N. Candy, "Bifid operators," *Math. Gazette*, vol. 38, p. 270; 1954.

[3] H. E. Stockman, "Pendulum parametric amplifier," *Am. J. Phys.*, vol. 28, pp. 506–507; May, 1960.

[4] Patent application of January 26, 1961, to U. S. Patent Office, Washington 25, D. C.

different schemes for synchronization were used, all but one of them free from mechanical switching. Thus, the source $i(t)$ in Fig. 1 may represent a tiny high-voltage battery (for example, a 600-volt battery), with the programming device consisting of an electronic switch, controlled capacitively, or otherwise, by the rotating vanes. For a slowly rotating motor, operated at 600 volts, the current drain is less than 1 μa, and the required power of the order of 100 μw. Since there are practically no ri^2-losses, the motor has high efficiency. The simplest programming system is obtained when vdC/dt yields a double-valued force function, sufficient all by itself to accomplish rotation.

Just as a conventional electric motor operates from a lead-cell storage battery, this new motor operates from a capacitor battery (the old Leyden jar battery). The experimental motor runs for several minutes from a 6-μf capacitor, initially charged to 800 volts. This type of motor can be made to produce considerable torque, however, since the air-space capacitance per unit volume can be given a high value and since air dielectric is only a first consideration. For space vehicles, a motor for high voltage and low current may fit available atomic and radiation sources better than conventional motors. If energy storage is not required, the motor may be ac-operated. The freedom from brushes should make this type of lightweight motor reliable and maintenance-free.

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The Doppler Effect and Inertial Systems*

If an artificial satellite S transmits a radio signal of frequency $f \gg f_c$, where f_c is the maximum plasma frequency of the ionosphere, an observer P on the earth records a frequency f'. Due to changes of the radial velocity between S and P, the received frequency f' varies in time as illustrated by the Doppler-shift curve of Fig. 1. For linear orbits of S and P, the time derivative of the Doppler-shift curve is described by the identity

$$\lambda(\dot{f}') = -\frac{v_{\mathrm{rel}}^2}{R}\sin^2\psi, \qquad (1)$$

where $\lambda = c/f$ is the wavelength of the emitted signal in free space, v_{rel} is the relative velocity between S and P, R is the range, and $\psi = \measuredangle v_{\mathrm{rel}}, R$. From (1) follows the minimum-range equation[1] ($\psi = \pi/2$, $R = R_m$)

$$\lambda(\dot{f}')_{\max} = -\frac{v_{\mathrm{rel}}^2}{R_m}, \qquad (2)$$

where $(\dot{f}')_{\max} = tg\alpha_{\max}$. Fig. 1 is the maximum rate of change of the observed Doppler frequency shift and R_m is the minimum range. With (2) one can evaluate R_m from the measurement of $(\dot{f}')_{\max}$ and from the knowledge of λ and v_{rel}. Eqs. (1) and (2) represent identities which are valid for linear orbits of S and P.

For circular orbits of S, with an observer P assumed at rest relative to this orbit, the minimum-range equation has been previously derived.[2] For the general case of curved orbits of S and P (Fig. 2), one obtains the minimum-range equation

$$\lambda(\dot{f}')_{f'=f} = -\frac{v_{\mathrm{rel}}^2}{R_m} + \frac{v_s^2}{\rho_s}\cos\alpha - \frac{v_p^2}{\rho_p}\cos\beta + \dot{v}_s\cos\gamma - \dot{v}_p\cos\delta, \qquad (3)$$

where v_s is the orbital speed of S and ρ_s is the radius of the orbit of S, v_p is the orbital speed of P and ρ_p is the radius of the orbit of P.

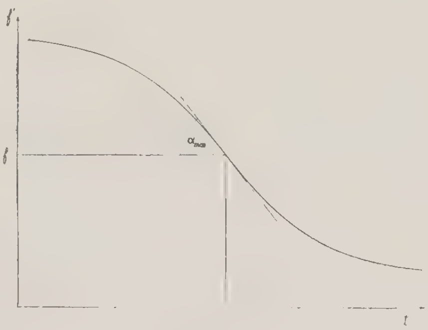

Fig. 1.

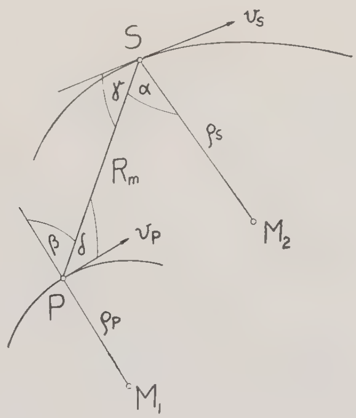

Fig. 2.

The second and third term in (3) represent the components of the radial accelerations of S and P in the direction of R_m with the signs corresponding to the angles α, $\beta < \pi/2$ as shown in Fig. 2. The fourth and fifth terms are, respectively, the components of the tangential accelerations of S and P in the direction of R_m for the general case of curved orbits with the signs corresponding to γ, $\theta < \pi/2$. For circular orbits, $\dot{v}_s = \dot{v}_p = 0$. If the orbits of S and P lie in the equatorial plane of the earth and one obtains, *e.g.*, $R_m = 1000$ km from the identity stated in (2), the second term of (3) increases R_m by 15 per cent while the third term of (3) reduces R_m by 0.1 per cent.

A comparison of (2) and (3) discloses that for curved orbits the orbital speeds v_s and v_p, although contained in v_{rel} according to the equation $v_{\mathrm{rel}}^2 = v_s^2 + v_p^2 - 2v_sv_p\cos\epsilon$, appear separately in (3). For $\dot{v}_s = \dot{v}_p = 0$ and $\rho_s = \rho_p = \infty$, (3) reduces to (2). Eq. (2) is valid for inertial systems. Those are systems without acceleration. For accelerated systems (curved orbits), the relative velocity v_{rel} describes inadequately the space-time continuum. Therefore, these nonrelativistic considerations illustrate the concepts of uniform and nonuniform relative motion which are known to underlie the special and the general theory of relativity, respectively.

If v_{rel} alone fulfills the identity of a space-time continuum (2), the special theory of relativity applies; the general theory of relativity is necessary, however, if v_s and v_p appear individually in the identity of a space-time continuum (3). Doppler studies of orbiting satellites are principally subject to the general theory because, for the description of the space-time continuum, v_{rel} and the concept of inertial systems as contained in (2) do not suffice. If they do it is only approximative.

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* Received by the IRE, January 30, 1961.
[1] W. Priester and G. Hergenhahn, "Bahnbestimmung von Erdsatelliten aus Doppler-Effekt Messungen," Westdeutscher Verlag, Köln und Opladen, West Germany; 1958. See especially p. 11.
[2] K. Toman, "The minimum-range equation and the maximum Doppler-frequency shift for satellites," Proc. IRE (Correspondence), vol. 48, pp. 1339–1340; July, 1960.

Envelope Probability as a Function of E/N_0*

It has been known for some years[1] that: 1) the parameter that determines probability of detection of a signal in additive noise is the ratio of signal energy to noise power per unit bandwidth; and 2) the optimum receiver uses a matched filter (or correlator), which generally does not produce a constant-amplitude sine wave at its output. Nevertheless, almost all (if not all) of the published analyses of detection probability emphasize signal-to-noise power ratio, and use as a model a constant-amplitude sine wave in noise at the input to the envelope detector. It is shown below that the probability density of the envelope of the output of a matched filter can easily be derived in terms of the ratio of signal energy to noise power per unit bandwidth. As a consequence, it is recommended that the use of "noise bandwidth," 3-db bandwidth, and 3-db pulse duration be restricted to situations in which they are really necessary.

The desired relationship can easily be derived with an error of 3 db as follows: Rice's[2] equation 3.10–11 for the probability density of the envelope R of a sine wave with constant amplitude P in additive narrow-

* Received by the IRE, February 6, 1961.
[1] J. L. Lawson and G. E. Uhlenbeck, "Threshold Signals," McGraw-Hill Book Co., Inc., New York, N. Y.; 1950.
[2] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. J.*, vol. 23, pp. 282–332, July, 1944; vol. 24, pp. 46–156, January, 1945.

band Gaussian noise with zero mean and variance ψ_0 gives:

$$p(R) = \frac{R}{\psi_0} \exp\left[-\frac{R^2+P^2}{2\psi_0}\right] I_0\left(\frac{RP}{\psi_0}\right). \quad (1)$$

Since papers on matched filters, such as Turin,[3] show that the peak signal-to-noise power ratio at the output of a matched filter is $2E/N_0$, where E is the input signal energy and N_0 is the input noise power per 1-cps band, one might be tempted to set $P^2/2\psi_0$ equal to $2E/N_0$ in (1). A suspicion that this is not correct could be aroused by an analysis of Manasse[4] that shows that envelope detection reduces the "effective E/N_0" by a factor of at least two. Closer reading of the papers on matched filters shows that envelope detection cannot be assumed in the argument leading to the conclusion that the power SNR is $2E/N_0$.

A direct, correct derivation of the desired relation can be obtained from Dugundji,[5] who shows very simply that, when the input to a (linear) filter with a transfer function $G(j2\pi f)$ is any signal $s(t)$ plus Gaussian noise with zero mean and power spectrum $W(f)$, the first probability density of the envelope R of the filter output is given by (1) above with $|z(t)|$ substituted for P, where:

$$z(t) = 2\int_0^\infty S(j2\pi f)G(j2\pi f)\exp[j2\pi ft]df \quad (2)$$

$$S(j2\pi f) = \int_{-\infty}^{\infty} s(t)\exp[-j2\pi ft]dt \quad (3)$$

$$\psi_0 = \int_{-\infty}^{\infty} |G(j2\pi f)|^2 W(f)df. \quad (4)$$

For simplicity, assume that the noise at the filter input is also white.[6] Let $W(f)=N_0/2$. The optimum (matched) filter can be defined[3] by:

$$G(j2\pi f) = kS^*(j2\pi f)\exp(-j2\pi ft_0) \quad (5)$$

where k is a real constant and $s(t)$ will produce a maximum filter output at $t=t_0$. Then

$$\psi_0 = \frac{N_0}{2}\int_{-\infty}^{\infty} |kS^*(j2\pi f)\exp(-j2\pi ft_0)|^2 df$$
$$= \frac{k^2N_0}{2}\int_{-\infty}^{\infty} |S(j2\pi f)|^2 df \quad (6)$$

and

$$z(t) = 2k\int_0^\infty |S(j2\pi f)|^2 \exp[j2\pi f(t-t_0)]df. \quad (7)$$

At $t=t_0$,

$$z(t_0) = 2k\int_0^\infty |S(j2\pi f)|^2 df$$
$$= k\int_{-\infty}^{\infty} |S(j2\pi f)|^2 df \quad (8)$$

and

$$\frac{|z(t_0)|^2}{2\psi_0} = \frac{1}{N_0}\int_{-\infty}^{\infty} |S(j2\pi f)|^2 df = E/N_0. \quad (9)$$

Thus, (1) above can be converted to apply to the general case of a matched filter followed by an envelope detector (which is optimum when the phase of the signal is unknown and has a constant probability density) by substituting E/N_0 for $P^2/2\psi_0$.

It is of course possible to interpret the above as proving that, for any matched-filter, envelope-detector receiver, there is an equivalent "Rice model" in which the signal has a constant power E/T for duration T, the noise power is N_0B, and $BT=1$. However, note that the definitions of B and T are completely arbitrary except for their product, so B and T need have no physical significance. Although B and T may be used for convenience in describing the receiver filter and the signal, their use is not required in the "Dugundji model," and this model is a more direct analog of the physical system than the "Rice model." Finally, the Dugundji approach shows directly the effects of using an imperfectly-matched filter, whereas the "Rice model" superficially implies that doubling B increases the required signal power by 3 db. Therefore, it is suggested that B and T be used only where their convenience really justifies their likely misuse.

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[3] G. L. Turin, "An introduction to matched filters," IRE TRANS. ON INFORMATION THEORY, vol. IT-6, pp. 311–329; June, 1960.

[4] R. Manasse, "The Performance of Post-Detection Integrators for the Detection of a Sine Wave in Noise at Low Input Signal-to-Noise Ratio," M.I.T. Lincoln Lab., Lexington, Mass., Group Rept. No. 32-26, ASTIA AD 236 234; November 28, 1956.

[5] J. Dugundji, "Envelopes and pre-envelopes of real waveforms," IRE TRANS. ON INFORMATION THEORY, vol. IT-4, pp. 53–57; March, 1958.

[6] Turin also derives a matched filter for nonwhite input noise.

8- and 11-Gc Nanosecond Carrier Pulses Produced by Harmonic Generation*

Oscillograms have been obtained showing carrier pulses of nanosecond duration and for which the RF carrier is phase locked with the envelope. These pulses have been obtained at carrier frequencies of both 8 Gc and 11 Gc and at pulse repetition rates of 10 and 20 Mc. Both a baseband electrical stroboscope[1] and a new band-pass electrical stroboscope were used to obtain these oscillograms.

The RF pulses are generated directly from the harmonics of the envelope frequency present in the sharp step that is found at the end of the recovery transient of a diffused silicon mesa computing diode.[2,3] The diode, a selected FD-100, is mounted across a waveguide and driven at the baseband repetition frequency. Only the harmonics higher in frequency than the cutoff frequency of the waveguide are propagated. In the present case, a 0.400×0.900-inch waveguide (WR-90) with a low-frequency cutoff of 6.56 Gc was used.

This microwave pulse generator circuitry is shown on the schematic of Fig. 1. From the secondary of the double-tuned output transformer, a high-level sine wave and dc from a self-biasing resistor are applied to the diode. The RF by-pass condenser C_1 is also the tuning capacitor for the secondary of the transformer.

In a test setup used to obtain the 8-Gc carrier pulses at a 20-Mc repetition frequency, a waveguide to coaxial transducer connected the microwave pulse generator of Fig. 1 directly to the coaxial input of the baseband electrical stroboscope.[4] Fig. 2(a) shows an oscillogram of this pulse. The horizontal sensitivity of the stroboscope is 5×10^{-10} seconds per large division. Since 4 phase-locked RF cycles are contained in

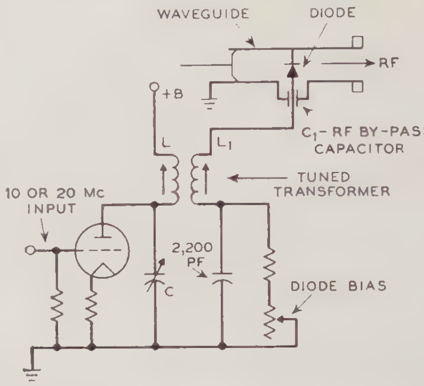

Fig. 1—Microwave pulse generator circuitry.

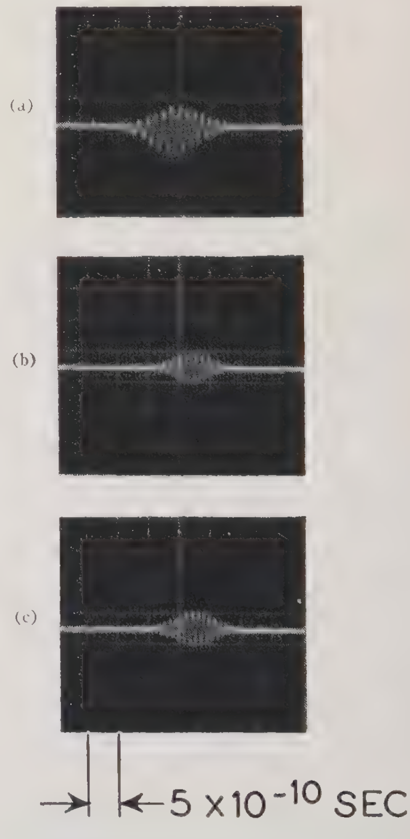

Fig. 2—Oscillograms showing phase-locked carrier cycles within the pulse envelope. (a) 8-Gc pulses at a 20-Mc repetition frequency. (b) 11-Gc pulses at a 10-Mc repetition frequency. (c) 11-Gc pulses at a 20-Mc repetition frequency.

* Received by the IRE, February 3, 1961.

[1] W. M. Goodall and A. F. Dietrich, "Fractional millimicrosecond electrical stroboscope," PROC. IRE, vol. 48, pp. 1591–1594; September, 1960.

[2] A. F. Boff, J. Moll and R. Shaw, "A new high-speed effect in solid state diodes," *Digest of Papers of 1960 Solid States Circuits Conf.*; pp. 50–51.

[3] A. F. Dietrich and W. M. Goodall, "Solid-state generator for 2×10^{-10} second pulses," PROC. IRE, vol. 48, pp. 791–792; April, 1960.

[4] Goodall and Dietrich, *op. cit.*; for block diagram of stroboscope, see Fig. 1, p. 1592.

one division, the time for an individual cycle is 1.25×10^{-10} seconds. This corresponds to a frequency of 8 Gc ($F=1/T$). The peak pulse power was measured as 0.17 mw.

Fig. 3 shows the block schematic of the band-pass electrical stroboscope used in the test setup to observe the 11-Gc carrier pulses at both a 10- and 20-Mc repetition frequency. The strobe pulse was generated by the same technique as in Fig. 1 but at a 100-cycle lower baseband frequency. This pulse is amplified by a traveling-wave tube which has a 3-db bandwidth of 2 Gc centered at 11.2 Gc. The amplified RF strobe pulse and a similarly amplified RF signal pulse are combined in a 3-db waveguide coupler and demodulated. The detected output, which is a low-frequency replica of the RF signal pulse, was applied to a low-frequency oscilloscope.

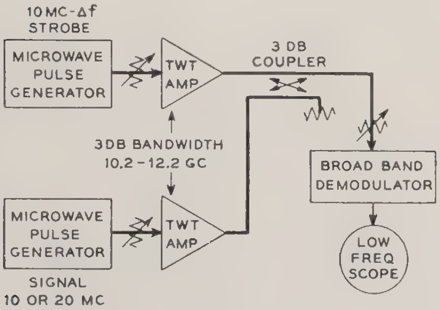

Fig. 3—Block schematic of test setup using band-pass electrical stroboscope.

Fig. 2(b) is an oscillogram of the 11-Gc, 10-Mc repetition-rate pulse with a peak power of 0.004 mw. Fig. 2(c) shows the 20-Mc repetition-rate pulse with a peak power of 0.011 mw. Horizontal sensitivity is the same as Fig. 2(a). In these two oscillograms there are 5.5 phase-locked RF cycles in the 5×10^{-10} seconds period. This corresponds to a frequency of 11 Gc. Using similar techniques, on April 7 oscillograms like Fig. 2 showing 56 Gc pulses at a 160 Mc repetition frequency were obtained.

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A Variable Dual Reactance Traveling-Wave Parametric Amplifier*

Traveling-wave-type parametric amplifiers have been realized which utilize transmission lines with one reactance being varied by a propagating wave. Tien and Suhl[1] have presented an analysis of parametric amplification in propagating circuits by using one time-varying distributed reactance (inductance) as the coupling element between two propagating media. Similar results can be achieved by employing two nonlinear distributed reactances (inductance and capacitance) to couple the two transmission lines. The mathematical model presented here, employing two variable distributed coupling reactances, shows that twice as much gain in nepers per unit length of line can be realized over an amplifier where only one coupling reactance is varied. Although the difference frequency between pump and signal is considered, the sum frequency and higher components are neglected, for they may be suppressed in the propagating structure by choosing the proper pump frequency or by designing a propagating structure such that the higher-order sidebands are outside the bandwidth of the circuit.

Thus, we shall consider the nondegenerate case ($\omega_1\neq\omega_2$) where waves of the signal frequency ω_1 on line No. 1 and waves of the difference or idler frequency ω_2 on line No. 2 are related as follows:

$$\omega_1+\omega_2=\omega. \tag{1}$$

These waves are coupled through two distributed reactances which vary in time at the pump frequency ω, and are spatially dependent upon the pump phase shift βz, where

$$\beta=\beta_1+\beta_2+\Delta\beta, \tag{2}$$

and z is the longitudinal direction of propagation. That is, the inductive and capacitive coupling reactances between these two lines may vary as follows:

$$L(z,t)=1/2L\left[e^{+j(\omega t-\beta z)}+e^{-j(\omega t-\beta z)}\right],$$

$$C(z,t)=1/2C\left[e^{+j(\omega t-\beta z)}+e^{-j(\omega t-\beta z)}\right],$$

when this coupling network is energized by a traveling-wave of frequency ω and phase constant β. When $\Delta\beta\neq0$, the system is no longer dispersionless, and it will be shown that the gain of the amplifier is reduced when $\Delta\beta$ deviates from zero. The two pairs of transmission line equations, neglecting series resistance and shunt conductance, are written as

$$\frac{\partial V_1(z,t)}{\partial z}=-L_1\frac{\partial I_1(z,t)}{\partial t}-\frac{\partial}{\partial t}\left[L(z,t)I_2(z,t)\right],$$

$$\frac{\partial I_1(z,t)}{\partial z}=-C_1\frac{\partial V_1(z,t)}{\partial t}-\frac{\partial}{\partial t}\left[C(z,t)V_2(z,t)\right]; \tag{3}$$

$$\frac{\partial V_2(z,t)}{\partial z}=-L_2\frac{\partial I_2(z,t)}{\partial t}-\frac{\partial}{\partial t}\left[L(z,t)I_1(z,t)\right],$$

$$\frac{\partial I_2(z,t)}{\partial z}=-C_2\frac{\partial V_2(z,t)}{\partial t}-\frac{\partial}{\partial t}\left[C(z,t)V_1(z,t)\right], \tag{4}$$

where V and I are, respectively, the voltage and current along a line. These two pairs of simultaneous equations can be readily employed to solve for the gain factor, α, of $I_1(z,t)$. To simplify the equations, let $\eta(\eta\ll1.0)$ be the ratio of the variable to the fixed reactance whether it be inductive or capacitive, assuming both to vary the same amount for mathematical simplicity. Thus, upon solving (3) and (4) for the gain factor α of the signal current $I_1(z,t)$, we have

$$\alpha=\tfrac{1}{2}(\eta^2\beta_1\beta_2-\Delta\beta^2)^{1/2}\text{ nepers per unit length.}$$

The corresponding term for a single variable coupling reactance (either inductance or capacitance) traveling-wave parametric amplifier is

$$\alpha=\tfrac{1}{4}(\eta^2\beta_1\beta_2-4\Delta\beta^2)^{1/2}\text{ nepers per unit length.}$$

Therefore, the gain in nepers per unit length α is doubled by employing two variable coupling reactances rather than one when $\Delta\beta=0$, *i.e.*, when the system is dispersionless. This condition exhibits optimum gain. As shown in Fig. 1 with $\Delta\beta\neq0$, the gain of the variable dual reactance amplifier decreases much less rapidly with increasing dispersion than does the single variable reactance device. Also, the dual amplifier can operate with considerable dispersion before the gain in nepers per unit length drops to a value equal to that of a nondispersive single-reactance amplifier. Thus, the dual device can be made to perform over a much greater gain-dispersion area than the conventional amplifier, provided the amplifier can be made stable. Finally, the dual amplifier would exhibit a better noise figure than the single amplifier if both devices exhibited approximately the same loss per unit length of line (with the pump off), since the dual device possesses more gain.

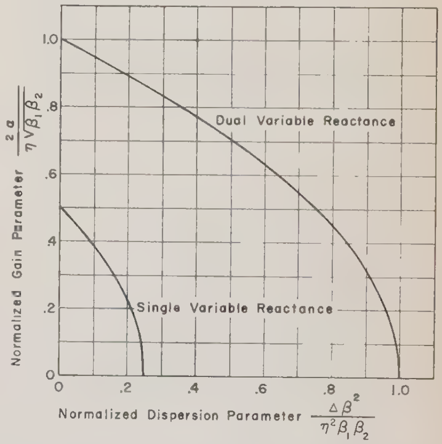

Fig. 1—Normalized gain dispersion characteristics for single and dual variable reactance traveling-wave parametric amplifiers.

This type of distributed parametric amplifier could possibly be realized with two transmission lines, in close proximity, immersed in a continuous medium consisting of both ferroelectric and ferromagnetic material, with the time-varying coupling capacitance and inductance being supplied by the nonlinear characteristics of each material, respectively. However, present-day materials would have to be greatly improved for such an amplifier to operate effectively. The entire structure might then be surrounded by some form of guided wave structure which would transmit the energizing pump wave. It should be noted that the distributed parametric amplifier is a relatively broad-band device and since the gain per unit length is proportional to the product of signal and idler phase constants, broad-band, low loss, slow-wave-propagating structures should be employed for high-gain and low-noise performance.

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* Received by the IRE, January 25, 1961; revised manuscript received, February 7, 1961.

[1] P. K. Tein and H. Suhl, "A traveling-wave ferromagnetic amplifier," Proc. IRE, vol. 46, pp. 700–706; April, 1958.

Parametric-Excited Resonator Using Junction Transistor*

In recent correspondence, the subject of the junction transistor as a three-terminal capacitor has been discussed.[1-3] A note by Giacoletto[3] describes an interesting phenomenon—the capacitive interaction between emitter and collector junctions of a transistor in the cutoff state.

The author discovered and utilized this phenomenon independently in early 1959, while working on the parametric-excited resonator using semiconductor junction devices, and reported it in his M.S. thesis (August, 1959), on which this note is based.

Fig. 1 shows a typical variation of the collector-junction capacitance due to the emitter-bias voltage for a medium frequency ($f_\alpha = 8$ Mc) alloy-junction transistor (2N123). As seen in the graph, with both junctions reverse-biased the capacitance of one junction can be controlled by the bias voltage at the other junction. This phenomenon may be explained qualitatively by the minority carrier distribution within the base region at different emitter-bias voltages (V_{eb}), as shown in Fig. 2(c). Thus, changing V_{eb} from reverse to forward bias causes the width of the emitter-junction depletion layer to vary due to field variation. At the same time, the width of the collector-junction layer is also affected. It is believed that it is mainly due to small changes in the diffusional transfer of thermally-generated minority carriers from the emitter junction to the collector junction. The equivalent circuit in this state is shown in Fig. 2(b). Since the transistor is in the cutoff state, both G_e and G_c are relatively high, and normal minority carrier injection, diffusion, and collection are not taking place.

Although variation of the collector-junction capacity due to the emitter-bias voltage changes in the cutoff state is small, this has some practical significance. With present transistors, which were not designed to be used as variable capacitors, we may not be able to excite the collector-resonance circuit by applying the pump voltage to the emitter, but we are certainly able to lock the phase of the collector oscillation signal (parametrically excited), by applying the phase-lock signal to the emitter junction. Utilizing the above characteristics, a successful parametric-excited resonator was built. A schematic of the experimental circuit is shown in Fig. 3(a). Collector circuit is essentially the parametric-excited resonance circuit tuned to the second subharmonic by collector dc bias-voltage V_{cb}, while the collector-junction capacity is being pumped by V_p at a frequency of 40 Mc. The phase-lock signal (20 Mc) is coupled to the emitter, and V_{eb} biases the emitter junction to the proper operating point.

* Received by the IRE, February 23, 1961. This note is based on "The Parametric Excited Resonator Using Semiconductor Devices," a thesis submitted by the author in partial fulfillment of the requirements for the M.S. degree in Elec. Engrg. at Mass. Inst. Tech., Cambridge, August, 1959.

[1] J. F. Gibbons and G. L. Pearson, "*P-N-P* variable capacitance diodes," Proc. IRE, vol. 48, pp. 253–255; February, 1960.

[2] J. M. Early, "*P-N-P* variable capacitance diode theory," Proc. IRE, vol. 48, pp. 1905–1906; November, 1960.

[3] L. J. Giacoletto, "Three-terminal variable capacitance semiconductor device," Proc. IRE, vol. 49, pp. 510–511; February, 1961.

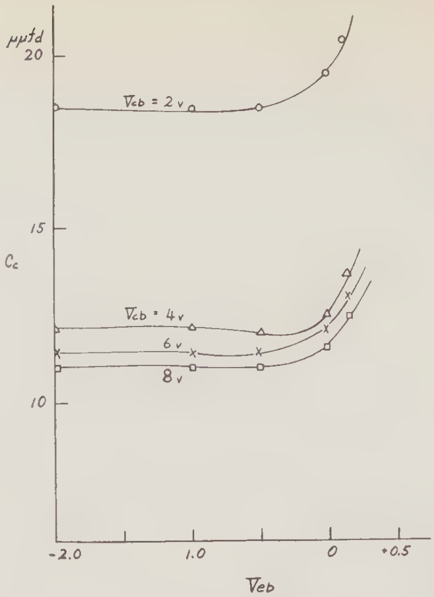
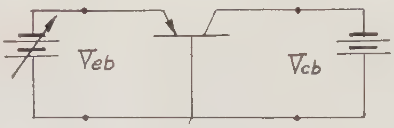
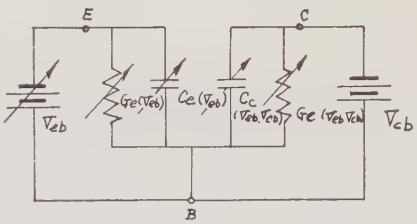


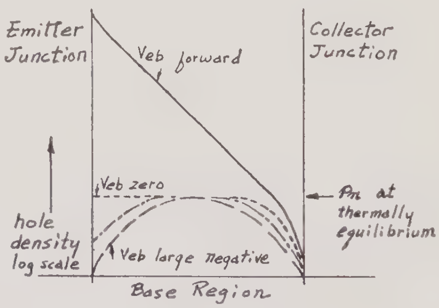
Fig. 1—Collector-junction capacity vs emitter-bias voltage.



(a)



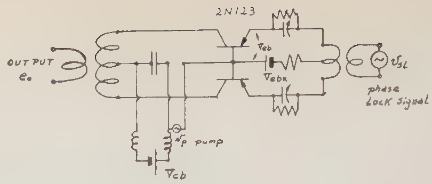
(b)



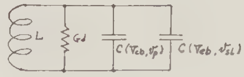
(c)

Fig. 2—(a) Reverse-biased junction transistor. (b) Equivalent circuit. (c) Minority carrier distribution within base region.

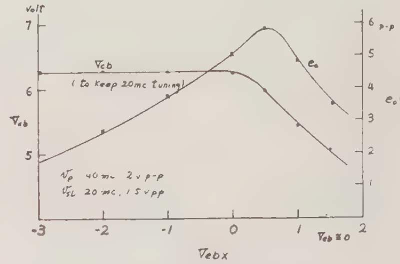
As seen in Fig. 1, in the vicinity of zero emitter bias, the capacitive interaction becomes most significant. But if the emitter should swing into the forward-conduction region, the transistor will become active (normal transistor action), and a relatively large dc current will flow through the collector circuit, damping the oscillation. The signal frequency (20 Mc) current ($\alpha \cdot ie$) due to normal transistor action is negligible because the signal frequency greatly exceeds the α-cutoff frequency of the transistor.



(a)



(b)



(c)

Fig. 3—Transistor parametric resonator: (a) experimental circuit; (b) equivalent circuit; (c) experimental result.

In Fig. 3(c), output voltage vs emitter dc bias-voltage for a constant drive was plotted. In the same graph, collector-detuning voltage was plotted against the emitter dc bias-voltage for a constant frequency of resonance.

The equivalent circuit of this parametric-excited resonator is shown in Fig. 3(b), and the diode parametric-circuit analysis which was derived by the author or that which was presented by Hilibrand and Beam[4] can be adopted with slight modifications.

The main advantages of this circuit as compared with the conventional diode parametric circuit are: 1) better isolation between input and output due to the use of a three-terminal device; 2) more important, the ability to switch the phase of the output signal without discontinuing the pump signal. The latter fact may be a very significant achievement in the application of the parametric circuit to high-speed phase switching. Because of this, the so-called three-phase pumping system required in the conventional (diode or magnetic core) parametric circuit may be eliminated.

It is well known that the limitations of an ordinary junction transistor in high-frequency applications are mainly due to α-cutoff (diffusion process) and the collector-junction capacity. Utilizing this junction capacity, once regarded as the troublesome factor, in a parametric-excited resonator, we are able to obtain a relatively high gain in the frequency range far beyond the α-cutoff frequency of the junction transistor.

The author is greatly indebted to Prof. Kyhl of MIT for his guidance throughout this work.

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[4] J. Hilibrand and W. R. Beam, "Semiconductor diodes in parametric subharmonic generator," *RCA Rev.*, pp. 221–253; June, 1959.

Noise Performance and Stability of a Hybrid-Coupled Tunnel Diode Amplifier*

An interesting hybrid-coupled tunnel diode amplifier has been described recently[1,2] which achieves very wide bandwidth, reasonable gain, low noise, and unconditional stability at the input. Since the performance data supplied by Sie[1] indicate favorable features only, it should be interesting also to recognize facts which limit the actual improvement of the noise performance of a system by using the hybrid-coupled tunnel diode amplifier in the front end of a receiver. This will be explained in the following.

Assume that a generator with an internal conductance G_g, a load G_L, and two equal tunnel diode amplifiers reflecting negative conductance $-G_D$ between the terminals, be properly connected to the hybrid of uniform characteristic conductance G_0. The power reflection coefficient at the hybrid terminals connected to the generator, the load, and the two equal amplifiers with tuned out susceptances are

$$\rho_g = \left(\frac{G_g - G_0}{G_g + G_0}\right)^2 \qquad (1a)$$

$$\rho_L = \left(\frac{G_L - G_0}{G_L + G_0}\right)^2 \qquad (1b)$$

$$\rho_D = \left(\frac{-G_D + G_1 - G_0}{-G_D + G_1 + G_0}\right)^2, \qquad (1c)$$

with G_1 = circuit loss of each tunnel diode amplifier reflected between the hybrid connecting points.

The effective gain of the hybrid amplifier when terminated into the load $G_L = G_0$ is equal to the power reflection at the tunnel diode amplifier terminations:

$$\gamma = \rho_D \qquad (2)$$

and the available gain is

$$\gamma_{av} = \gamma(1 - \rho_g) = \rho_D(1 - \rho_g). \qquad (3)$$

For the calculation of the noise figure, $G_L = G_0$ is assumed corresponding to $\rho_L = 0$, since this is the interesting condition for unconditional input stability. For the total noise output into $G_L = G_0$ results

$$N_0 = \gamma(1-\rho_g)kT_gB + \gamma^2\rho_g kT_LB + \frac{4kT_DBG_D'G_0}{(G_0 + G_1 - G_D)^2}(1 + \rho_g\gamma) + \frac{4kT_1BG_1G_0}{(G_0 + G_1 - G_D)^2}(1 + \rho_g\gamma). \qquad (4)$$

(T_g, T_L, T_D, T_1 = noise temperatures of generator, load, tunnel diodes and losses.) The first term in this equation is the amplified available generator noise power. The second term is originated by the noise power kT_LB entering the amplifier from the output termination G_L, of which the portion $\gamma^2\rho_g kT_LB$ is delivered back to G_L. The third and fourth terms express the output noise originating from the tunnel diode amplifiers. First, the noise power delivered by each tunnel diode amplifier into the hybrid may be expressed by

$$N_D + N_1 = \frac{4kT_DBG_D'G_0}{(G_0 + G_1 - G_D)^2} + \frac{4kT_1BG_1G_0}{(G_0 + G_1 - G_D)^2}, \qquad (5)$$

with G_D' = effective noise generating conductance of the lossy tunnel diode. Half of this power will go into G_L directly, the other part is directed towards the generator. There a portion of it is reflected back into the hybrid, and subsequently it will appear in the load G_L, multiplied by the factor $\rho_g\gamma$.[3] The same noise power is contributed by the other tunnel diode amplifier. The noise figure that results is:

$$F = \frac{N_0}{\gamma(1-\rho_g)kT_gB} = 1 + \gamma\frac{\rho_g}{1-\rho_g}\frac{T_L}{T_g} + \frac{4G_D'G_0(1+\rho_g\gamma)}{(1-\rho_g)(G_0 + G_D - G_1)^2}\cdot\frac{T_D}{T_g} + \frac{4G_1G_0(1+\rho_g\gamma)}{(1-\rho_g)(G_0 + G_D - G_1)^2}\cdot\frac{T_1}{T_g}. \qquad (6)$$

Under ideal conditions, there is no reflection at the input, and any losses in the circuit and the tunnel diodes can be neglected. Then the noise figure becomes:

$$F = 1 + \frac{4G_D/G_0}{(1 + G_D/G_0)^2}\cdot\frac{g_e}{g_D}\cdot\frac{T_D}{T_g}, \qquad (7)$$

with $G_D = \eta g_D$, $-g_D$ = negative conductance of the diodes, $g_e = eI_0/2kT_D$, η = transformation factor.

The noise figure is directly related to the gain. From (1c) and (2), one obtains

$$G_D/G_0 = 1 - \frac{2}{\sqrt{\gamma} + 1} \qquad (8)$$

and consequently

$$F = 1 + \left(1 - \frac{1}{\gamma}\right)\frac{g_e}{g_D}\frac{T_D}{T_g} = 1 + \left(1 - \frac{1}{\gamma_{av}}\right)\frac{g_e}{g_D}\frac{T_D}{T_g} = 1 + \left(1 - \frac{1}{\gamma_{av}}\right)\frac{eI_0}{2kT_g g_D}. \qquad (9)$$

This expression shows that for a given T_g, and diode parameters, the ideal noise figure of the hybrid amplifier depends only on the gain γ, and that the noise figure increases with the gain. The noise figure varies from 1 to a value asymptotically approaching $1 + eI_0/2kT_g g_D$ as the gain is varied from 0 db to large values. The hybrid amplifier shows an improvement by the factor $\gamma_{av} - 1/\gamma_{av}$ compared to the noise of a simple parallel connected tunnel diode amplifier, whose optimal noise figure is $F = 1 + eI_0/2kT_g g_D$ for infinite load impedance. The reason for this improvement is the elimination of the load noise contribution.

The question as to what extent a system is improved using a tunnel diode hybrid preamplifier can be answered as follows: Using (9), the noise figure of a system becomes

$$F_{\text{system}} = F_1 + \frac{F_2 - 1}{\gamma_1} = 1 + \frac{g_e}{g_D}\frac{T_D}{T_g} + \frac{F_2 - (1 + g_e/g_D T_D/T_g)}{\gamma_1}. \qquad (10)$$

From this the following conclusions are obvious: if the noise figure F_2 of the succeeding stage is greater than $1 + eI_0/2kT_g g_D$, the noise figure of the system can be reduced to this value for $\gamma_1 \to \infty$, ideally. However, if F_2 is smaller than $1 + eI_0/2kT_g g_D$, an improvement of the system is impossible.

Another remark shall concern the stability of the amplifier. If the amplifier is terminated by a load $G_L = G_0$, then with respect to a wave going into the input no wave can ever be reflected, which, at first, means unconditional input stability. However, a stability condition has to be met by the amplifier not only with respect to an incident wave, but also with respect to the varying impedance which appears at the *TD* amplifier terminals of the hybrid if the generator impedance is varied. This restricts the range of stability, in particular at high gain.

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[1] J. J. Sie, "Absolutely stable hybrid-coupled tunnel diode amplifier," Proc. IRE (Correspondence), vol. 48, p. 1321; July, 1960.

[2] J. J. Sie, "Correction to 'Absolutely stable hybrid-coupled tunnel diode amplifier'," Proc. IRE (Correspondence), vol. 48, p. 1783; October, 1960.

[3] No correlation between the two portions is assumed.

Millimeter-Wave Field-Displacement-Type Isolators with Short Ferrite Strips*

Generally, field-displacement-type isolators have more compact and simple structures than Faraday rotation-type isolators. At millimeter waves this is, however, not always true. For example, the isolator developed by Weiss and Dunn required a ferrite strip length of two inches for 5 mm wavelength.[1] Ayres developed an isolator with a ferrite for 4 mm wavelength with the relatively high magnetic flux density of 12 kilogauss.[2] These long thin ferrite strips break easily.

Recently, field-displacement-type isolators of extremely short ferrite strip (only 0.2 inches) with low magnetic field (less than 5 kilo-oersted) for 5 mm wavelength have been developed by the authors. The ferrite strip was made from a single-crystal-type sample LRR-1 supplied by A. O. Smith Corporation. The sample had an anisotropy field of 18.4 kilo-oersted and a line width of approximately 10 oersted. The sample was magnetized in the direction of easy magnetization.

* Received by the IRE, February 27, 1961. This research was supported by a research grant from A. O. Smith Corp. to Marquette University.

[1] M. T. Weiss and F. A. Dunn, "A 5-mm resonance isolator," IRE Trans. on Microwave Theory and Techniques, vol. MTT-6, p. 331; July, 1958.

[2] W. P. Ayres, "Millimeter-wave generation experiment utilizing ferrites," IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 62–64; January, 1959.

The ferrite strip of 0.209″×0.0325″ ×0.0117″ was mounted on a polystyrene strip of the same length. This was placed in RG-98/U waveguide as shown in Fig. 1. The external magnetic field was applied parallel to the crystal axis. The magnetic field was applied by an adjustable permanent magnet. This was adjusted for best performance at the operating frequency.

An example of experimentally obtained operating frequency characteristics of this device is shown in Fig. 2. When the operating frequency is changed, the applied magnetic flux density is changed as indicated in the figure. As is well known, the square of the gyromagnetic resonance frequency f_0 can be expressed as a constant coefficient second-order function of the applied magnetic field.[3] If B_0 represents the magnetic flux density in kilo-gauss measured at the air gap where the ferrite strip is mounted when the ferrite resonates, the following equation can be obtained from the experimental data:

$$f_0^2 = 30.6B_0^2 + 38.1B_0 + 2799 \ (kMc)^2.$$

This equation can help to estimate performance over a wide range of operating frequency and magnetic flux density.

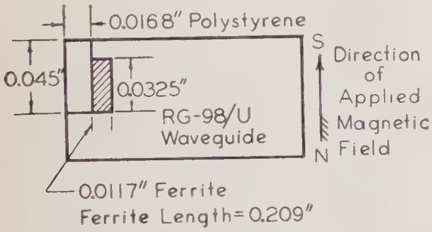

Fig. 1—Cross section of millimeter-wave field-displacement-type isolator.

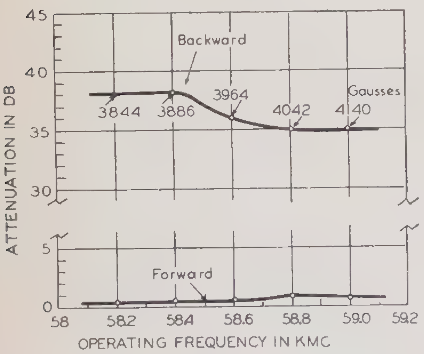

Fig. 2—Operating frequency characteristics of millimeter-wave field-displacement-type isolator.

The frequency bandwidth for 15 db isolation for a fixed magnet adjustment was 190 Mc. The forward attenuation was less than 1 db over the entire band. The frequency bandwidth can be increased by increasing the ferrite strip length. The backward attenuation increased faster than forward attenuation with increasing ferrite strip length.

Even though the mounting position and geometrical shape of the ferrite strip rather critically influenced the isolator characteristics, frequency characteristics of the figure of merit of this isolator $F \simeq (4\omega/\gamma\Delta H)^2$ were flat for wide range of operating frequencies.[3] The reason is that the line-width ΔH of this material increased linearly with operating frequency. The short strip makes the isolator compact and it makes production easy and inexpensive.

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[3] B. Lax, "Frequency and loss characteristics of microwaves ferrite devices," Proc. IRE, vol. 44, pp. 1368–1386; October, 1956.

Effect of Filament Magnetic Field on the Electron Beam from a Pierce Gun*

The effect of the magnetic field of a flat, spiral-wound, "noninductive" filament on the electron beam from a shielded Pierce electron gun has been measured by use of a beam tester that has recently been constructed at Cornell University, Ithaca, N. Y. A more detailed report describing the beam tester and other results of tests on dc electron beams will be published in the near future.[1] The beam is produced by a Pierce gun which normally operates at 5400 volts and has a perveance of 1.15×10^{-6}. The cathode diameter is 1.25 inches and the Brillouin beam diameter containing 90 per cent of the beam current is 0.176 inch. The effect of the magnetic field of the filament on the beam shape was determined by taking cross sections, similar to those shown in Fig. 1, along the axis of the beam for filament currents from −7.65 to +7.65 amperes. The two beam cross sections shown in Fig. 1 were taken with an x-y recorder. The horizontal and vertical coordinates are the x and y positions in the beam. The deflection above any horizontal line indicates the amount of current collected by a Faraday cage positioned behind a movable plate with a 0.010-inch hole in the center.

The instantaneous filament current was varied while keeping the cathode temperature constant by applying a sinusoidal 60-cps voltage to the filament and by pulsing the cathode at a rate of 60 cps with 16.7-μsec duration pulses. The phase of the cathode pulse relative to the filament voltage was then made variable from −90° to +90°. The curves in Fig. 2 show the maximum beam diameters containing 90 per cent and 95 per cent of the beam current as a function of the current through the filament. Note that the beam diameter obtained when the maximum filament current was present was over 13 per cent greater than that obtained when no filament current was present. In addition to this increase in diameter, a large amount of translaminar current has been noted at all times when the cathode is pulsed while filament current is flowing.

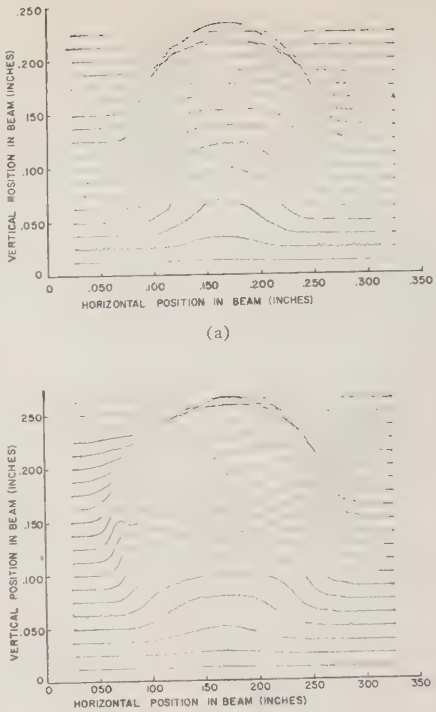

Fig. 1—Electron beam cross sections taken with an x-y recorder for (a) Brillouin flow conditions with no current through filament (less than 2 per cent scallop on beam), (b) Brillouin flow conditions with 7.65 amperes through filament.

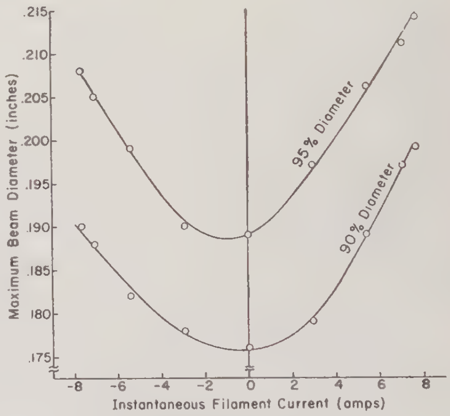

Fig. 2—Maximum beam diameter vs instantaneous filament current.

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* Received by the IRE, February 21, 1961.

[1] This work is part of a study of the RF characteristics of electron beams which is being conducted under the sponsorship of the Rome Air Dev. Center of the Air Res. and Dev. Command, USAF (R.F. Transmitter and High Power Tube Branch).

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He was with Westinghouse Electric Company from 1940 to 1945. In the latter part of 1945, he joined the Polytechnic Institute of Brooklyn, where he now holds the position of Research Professor and Acting Head of the Electrophysics Department. He is also a member of the Board of Directors of the Columbia Technical Corporation. His major work has been in the fields of microwave devices and network theory.

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He was associated with the Collins Radio Company during the summers of 1937 and 1938, and with the Bell Telephone Laboratories in 1940. From 1946 to 1956, he was associated with RCA Laboratories, Princeton, N. J., where he was first occupied with electron tube research and subsequently with transistor and semiconductor research. In connection with the latter work, he has received world-wide recognition for the development of a transistor parameter circuit which bears his name and for the first development of a nonlinear semiconductor capacitor with ultra-high-frequency properties which is now being exploited extensively for low-noise microwave parametric amplification.

From 1956 to 1960, he was Manager of the Electronics Department of the Ford Motor Company Scientific Laboratories, Dearborn, Mich. He is presently a professor in the Department of Electrical Engineering and the Division of Engineering Research within the College of Engineering, Michigan State University, East Lansing.

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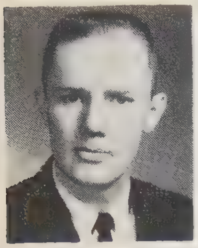

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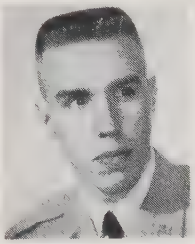

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❖

Philip M. Mostov was born in New York, N. Y., on August 2, 1921. He earned the B.S.E.E degree in 1944 from the College of the City of New York, the M.S.E.E. degree in 1947 from Columbia University, New York, N Y., and the Ph.D. degree in physics in 1953 from New York University.

Dr. Mostov has had extensive experience in physics, electrical engineering, and applied mathematics. During 1951–1954, he was associated with the Institute for Mathematics and Mechanics, New York University. During 1954–1956, he was with the Reactor Division of Combustion Engineering, Inc., New York, N. Y. During 1956–1958, he was connected with the Nuclear Development Corporation of America, consultants, in White Plains, N. Y. In 1958, he joined the Scientific Research Staff of Republic Aviation Corporation, Farmingdale, N. Y.; and in 1959, he helped found the Plasma Propulsion Laboratory at Republic. He is currently engaged in the study of plasma acceleration mechanisms suitable for propulsion, fusion injection and electromagnetic guns, as well as in the study of novel nuclear plasma energy conversion schemes. He also spearheads into crossed disciplines, creating and evaluating ideas and devices based upon plasma, MHD, fission and fusion technologies. He has taught both physics and electrical engineering.

P. M. Mostov

Dr. Mostov is a member of several professional and honorary societies.

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Joseph L. Neuringer was born in Brooklyn, N. Y., on January 16, 1922. He received the B.A. degree in mathematics and physics from Brooklyn College, N. Y., in 1943, the M.A. degree in physics from Columbia University, N. Y., in 1948, and the Ph.D. degree in physics from New York University in 1951. While at N.Y.U., he worked as a research assistant on Project Squid, and on various jet-propulsion research projects.

J. L. Neuringer

In 1951, he joined the Republic Aviation Corporation Farmingdale, N. Y., and participated in research and development work in aerothermodynamics, aeroelasticity and the dynamics of supersonic aircraft until 1955. From 1955, to 1956, he worked for Avco Research and Advanced Development Division, Wilmington, Mass., where he conducted research and development work on re-entry problems relative to the I.C.B.M. program. He rejoined Republic Aviation in 1956 and engaged in basic research in magnetohydrodynamics and plasma physics, and in 1959 was promoted to Chief of Theoretical Analysis in the Plasma Propulsion Laboratory; early in 1960, he was promoted to the position of Chief Scientist in that laboratory.

Dr. Neuringer is a member of the American Physical Society, the American Rocket Society, Sigma Xi, and Pi Mu Epsilon. He is a member of the Committee on Magnetohydrodynamics of the American Rocket Society and is also listed in "Who's Who in World Aviation" and the 1960 Edition of "American Men of Science."

Robert W. Olthuis (S'44–A'51–M'56), was born in Battle Creek, Mich., on October 11, 1921. He received the M.E. degree from Stevens Institute of Technology, Hoboken, N. J., in 1944. After military service, he received the M.S.E.E. and Ph.D. degrees from the University of Michigan, Ann Arbor, in 1947 and 1951, respectively. He joined the Sperry Gyroscope Company, Great Neck, N. Y., in 1950, where his work has been on microwave tubes and where he is now senior engineer in charge of a group doing cathode research.

R. W. Olthuis

Dr. Olthuis is a member of the American Physical Society, Tau Beta Pi, and Sigma Xi.

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Donald S. Rigney was born in Wilbur, Wash., on November 27, 1936. He received the A.B. and M.S. degrees in physics from San Diego State College, San Diego, Calif., in 1958 and 1959, respectively.

D. S. Rigney

From June, 1956, to September, 1957, he was associated with Convair, San Diego, working in the electromechanical packaging and electronic components test groups. In July, 1959, he joined the Plasma Propulsion Laboratory, Republic Aviation Corporation, Farmingdale, N. Y., where he has been involved in plasma acceleration studies, analysis of power subsystems for space applications, and network synthesis problems. He is currently analyzing the space-time history of the current discharge in a spherical theta-pinch, and studying the electromechanical interactions in the pinch process.

John A. Seeger (S'58–M'59) was born in Seattle, Wash., on June 24, 1929. He received the B.A. degree in physics from Pomona College, Claremont, Calif., in 1956, and the M.S.E.E. degree from the University of California, Berkeley, in 1958, his research having been done under the direction of Prof. Charles Süsskind in the field of microwave tubes.

J. A. Seeger

From 1958 to 1959, he worked at the U. S. Naval Ordnance Laboratory, Corona, Calif., where he did research in miniaturization of antennas. In 1959, he transferred to the Physics Division of Michelson Laboratory, China Lake, Calif., where he is presently a physicist with the Microwave Physics Branch. His current research is concerned with electrostatic focusing systems for microwave tubes.

Mr. Seeger is a member of the American Physical Society and Sigma Xi.

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D. C. Youla was born in Brooklyn, N. Y., on October 17, 1925. He received the B.E.E. degree from the College of the City of New York in January, 1947, and the M.S. degree from New York University, New York City, in June, 1950.

From 1947 to 1949 he was employed as an instructor in the Department of Electrical Engineering at C.C.N.Y. He attended the N.Y.U. Graduate School of Mathematics as a full-time student from 1948 to 1950, and for the next two years was at Fort Monmouth, N. J., and Brooklyn Naval Shipyard working on problems of UHF and microphonics. In 1952, he joined the communication group at the Jet Propulsion Laboratories, Pasadena, Calif., and participated in the design of antijam radio links for guided missiles. In 1955, he began his present association with the Microwave Research Institute of the Polytechnic Institute of Brooklyn where he engaged in the practical and theoretical study of codes for combating noise and improving efficiency. He is now a research associate professor of electrical engineering working actively on network synthesis problems, stability of time variable systems, solid-state devices and n-port filtering.

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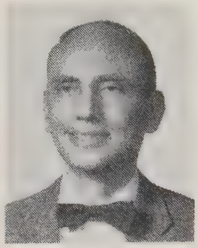
Glen Wade (S'51–A'55–SM'57) was born in Ogden, Utah, on March 19, 1921. He received the B.S. and M.S. degrees, both in electrical engineering from the University of Utah, Salt Lake City, in 1948 and 1949, respectively, and the Ph.D. degree from Stanford University, Stanford, Calif., in 1954. At Stanford, he was first a Sperry Fellow and then an RCA Fellow in electronics.



G. Wade

He served in the U. S. Navy in World War II as an electronics technician and later as an electronics officer. From 1949–1950, he was employed at the Naval Research Laboratory, Washington, D. C. In 1954, he was employed as a research associate by the General Electric Microwave Laboratory at Stanford University. Until recently, he has been an associate professor of electrical engineering at Stanford and a senior staff member of the Stanford Electronics Laboratories. He has also served as a consultant for the Zenith Radio Corporation, Philco Corporation, and Diamond Ordnance Fuze Laboratory. Early in 1960, he joined the Raytheon Company, Spencer Laboratory, Burlington, Mass., as an Associate Director of Engineering for General Research.

Dr. Wade received an Eta Kappa Nu Award in the "Outstanding Young Electrical Engineer" competition in 1955, and a National Electronics Conference Annual Award in 1959. He is a member of the American Physical Society, Phi Kappa Phi, Tau Beta Pi, Eta Kappa Nu and Sigma Xi.

Books

Control System Analysis and Synthesis, by John J. D'Azzo and Constantine H. Houpis

Published (1960) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 489 pages+14 index pages+xii pages+42 appendix pages+24 problem pages. Illus. $6\frac{1}{8}\times 9\frac{1}{4}$. \$13.50.

As a very lucid and concise exposition of control system principles, this book is ideally suited for a first course in feedback control at the senior or graduate level. Both as a teaching tool and an instrument of self-study, it is one of the best textbooks available.

A solid background is provided in the first five chapters, which cover the writing of differential equations of control elements, classical and Laplace transform analyses, and the expression of these equations in terms of transfer functions and block diagrams. The root locus and frequency response techniques are then developed and applied to numerous examples to show the effects of difference types of cascade and feedback compensation. A clear development of the Nyquist Criterion is given, as an introduction to the frequency response approach.

In addition to the basic material, there are chapters on ac carrier systems, complex multiloop control systems, nonlinearities, criteria for optimum response of control systems, and the use of analog computers in design. The chapter on nonlinearities is based upon the describing function, and gives a rather extensive description of its application. The exposition of all this material is very clear and is well supplemented by a large number of examples.

On the other hand, the book reflects what is seen by this reviewer to be a widespread failure in the field to grasp certain essentials of feedback control dynamics. (In the remarks to follow it is not to be inferred that the authors of this text are being singled out for criticism.)

An example of this failing is the definition of servo bandwidth that D'Azzo and Houpis employ for frequency response design. They define servo bandwidth in terms of the frequency of peak magnitude of the closed-loop frequency response, called the resonant frequency; and they claim that the higher the resonant frequency the faster the response. However, it is the phase curve rather than the magnitude curve that describes the speed of response and dynamic accuracy of a servo, because phase lag represents the time delay, and it is time delay between input and output that produces dynamic error.

The practice of defining servo bandwidth in terms of the magnitude response was borrowed incorrectly from the communications field. The magnitude response describes waveform fidelity and the steepness of the front of the step response, which are important in communications. However, the phase curve describes the actual time delay between the application of a step input and the rise of the step response, which is important in control but usually is not important in communications. For example, consider the step response of a pure delay line. The front of the response is very steep (which is reflected by the wide bandwidth of the magnitude curve), but the time delay between input and response is determined by the phase curve.

This distinction between magnitude and phase in the specification of bandwidth is not academic! The parameters of a servo can readily be altered so as to change its resonant frequency by an order of magnitude and still keep a constant value for the frequency of one radian of phase lag. In fact, Fig. 10–35 of the book gives an example of a servo with a resonant frequency three times that of the frequency of one radian of phase lag, whereas in other examples these frequencies are nearly equal. The frequency of one radian of phase lag is always approximately equal to the open-loop gain-crossover frequency; and either of these represents a good bandwidth criterion, because its reciprocal is approximately equal to 1) the time for the step response to rise to 63 per cent and 2) the maximum time delay in the ramp response between input and output.

There is also widespread confusion in the field concerning many other basic aspects of feedback control. For example, when does an integral network improve performance, and what are its limitations? When is a system in the steady state, so that error coefficients can be applied, and how do we define performance prior to that point? Thus, even though this book provides an excellent exposition of conventional feedback control material, it still fails to answer many fundamental questions which are of great practical importance in feedback control design.

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Complex Variables and the Laplace Transform for Engineers, by Wilbur R. LePage

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 464 pages+5 index pages+xvii pages+2 bibliography pages+4 appendix pages. Illus. $6\frac{1}{4}\times 9\frac{1}{4}$. \$12.50.

In the preface the author states that the book is written "for the serious student, probably at the graduate level, who is interested in obtaining an understanding of the theory of Fourier and Laplace transforms, together with the basic theory of a complex variable, without which the transform theory cannot be understood." The book is indeed well suited for classroom instruction as well as self-study.

In the main, the book is divided into two parts, as indicated by its title. The first part, on the theory of a complex variable, starts out with an introductory chapter giving an outline of the structure of system analysis and explaining some basic mathematical and engineering terms. Chapter 2 treats the foundation of the theory of a complex variable centered around the Cauchy-Riemann equations. The next three chapters, which treat conformal mapping, complex integration, and infinite series, lead up to an important chapter on multivalued functions. The first part of the book ends with a "mixed" chapter on diverse theorems which are of importance in network theory.

The second part, on the Fourier and Laplace transform theory and some of its applications in engineering, starts out with the basic Chapter 8 on real integrals, in which the main stress is put on a discussion of improper integrals. The Riemann integral representation is used. Then three important chapters follow on the Fourier integral, the Laplace integral, and convolution ("Faltung") integrals. Both the one-sided and the two-sided Laplace integral, the latter well known from the work by van der Pol and Bremmer, are discussed. After another "mixed" chapter on some additional properties of the Laplace integral, the second part of the book ends with four chapters (13–16) on the application of transform theory to the solution of ordinary linear integrodifferential equations with constant coefficients, impulse functions, periodic functions, and the Z transform.

The task of writing a mathematics book for engineers is not an easy one. The author, a well-known scientist and teacher, has done a good job in finding a balance between the dry, rigorous presentation in mathematics books and the often too heuristic presentation in engineering books. The mathematical ideas are lucidly presented in the easily-read text with its well-chosen illustrative figures.

A minor detail that may disturb the reader is the spelling of the name of the French mathematician l'Hospital (1661–1704), the inventor of the famous l'Hospital's theorem in calculus (spelled "Lhopital"). To the reviewer's knowledge, no book on complex variables for engineers has yet treated linear fractional transformations by the isometric circle method. Also, when will the Lebesque and Stieltjes integral representations find their rightful places in the mathematics-for-engineers textbook literature?

The bibliography contains many good references. One important oldtimer is missing, the "Funktionentheorie und ihre Anwendung in der Technik," by R. Rothe, F. Ollendorff, and K. Pohlhausen (Julius Springer, Berlin, 1931; translated by A. Herzenberg under the title, "Theory of Functions as Applied to Engineering Problems," The Technology Press, Massachusetts Institute of Technology, Cambridge, Mass., 1933).

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Relativistic Electron Theory, by M. E. Rose

Published (1961) by John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. 272 pages+7 index pages+xiii pages+bibliography by chapter+19 appendix pages+1 page general references. Illus. 6×9¼. $9.50.

This is a comprehensive account of the single particle Dirac theory, that is to say, the theory of all spin one-half particles (fermions) which are lighter than a nucleon, and which are not subject to strong couplings. The word "electron" in the title stands for the ordinary electron, the mμ meson, the neutrino, and their antiparticles as well. This treatment of the single particle description of relativistic quantum mechanics is intended primarily for physicists who have had a thorough grounding in the general principles and methods of nonrelativistic quantum mechanics. It is suitable for use as a graduate text, as well as a reference for workers in the field.

The book opens with a concise review of the nonrelativistic quantum theory of a spinning electron. This is followed by a thoroughgoing account of the relativistic quantum mechanics of a free particle, of a particle in an electromagnetic field, and of a particle in a central field. Problems which can be solved exactly, as well as some which can be solved only approximately, are carefully and lucidly discussed. There is an up-to-date account of neutrino theory and of the non-conservation of parity. Included in this book are a number of useful problems, an adequate set of references to the original literature, and some worthwhile appendixes on items such as the Lorentz transformations.

The author, a chief physicist at the Oak Ridge National Laboratory, has now written three excellent monographs on mathematical physics. The first two, "Multipole Fields" and "Elementary Theory of Angular Momentum," and the present contribution as well, are all models of clear and elegant mathematical exposition. All three represent very welcome additions to the literature. The author is to be commended for his very successful attempts to present very difficult material in a clear, well-organized, and physically interesting manner. His present contribution is highly recommended to all scientists who have the mathematical and physical background necessary to appreciate a work of this kind.

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Electromagnetic Fields and Waves, by Robert V. Langmuir

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 208 pages+3 index pages+ix pages+2 bibliography pages+14 appendix pages. Illus. 6×9¼. $9.75.

This is a book which should find wide use as a text for a senior course for electrical engineering majors. Although many of the sections throughout the book are similar to those traditionally included in most senior-level texts on electromagnetics, many topics which have been normally included in more advanced texts are introduced here. In particular, there are several examples involving line charges within conducting boundaries. The first two chapters are copiously supplied with examples carefully chosen so as to illustrate application of the laws and concepts under consideration. In general the examples extend the scope of the text as well. The problems at the end of each chapter represent exercises of varying difficulty, and it is in many of these that the student encounters subjects not treated in the text proper but which he is able to handle.

The small size of the book does not imply that the scope of the text is limited, but attests to the efficient use of the pages by the author. Each chapter is introduced by a survey of its contents which is useful in a text. The chapter content is as follows: Chapter I—The subject of electrostatics is introduced in the traditional manner. Chapter II—The divergence theorem is derived, after which solutions for Laplace's equation are discussed. Superposition of solutions and series solutions involving Gauss' law are carried out in detail. Examples are presented for rectangular, spherical, and cylindrical coordinate systems. Chapter III—Dielectric phenomena are considered for polarized materials. Problems involving dielectrics are solved, some by using the imaging techniques. Chapter IV—Magnetism is discussed in greater detail than in most books of this level. The treatment is carried over into Chapter V in which magnetic materials, including magnet design, are considered. Chapter VI—Maxwell's equations are introduced, inductance and energy storage are described and the wave equation is obtained. Chapter VII—The wave equation is specialized to the plane wave case. Poynting's theorem is derived. Chapter VIII—Rectangular waveguides and wave velocities are considered with extension to parallel plane guides. Chapter IX—In this chapter, skin effects and losses in waveguides are discussed, as is the eddy current in pulse transformer laminations. Chapter X—Reversing the usual order, it is in this chapter that TEM waves are presented. Chapter XI—The solution of the wave equation in spherical coordinates leads to radiation from simple antennas. Chapter XII—Cavities and cylindrical waveguide modes are considered. Chapter XIII—This chapter contains some topics in wave propagation. The reflection and refraction of plane waves, properties of dielectrics and conductors at various frequencies, surface guided waves, ionospheric propagation and dielectric rod waveguides are presented in enough detail so as to familiarize the student with the topics.

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Analytical Techniques for Non-Linear Control Systems, by John C. West

Published (1961) by D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. 186 pages+2 index pages+xii pages+4 bibliography pages+18 appendix pages. Illus. 5½×8½. $5.75.

The past five years have witnessed, not only rapidly expanding research interest in nonlinear control systems, but also the appearance of numerous textbook treatments of the subject. In the present book, Professor West has elected to focus attention on three basic topics—phase-plane analysis, describing functions, and the response to random signals—and place the major emphasis on the presentation of the fundamental concepts underlying each area.

Thus, the emphasis throughout the book is on an introduction to these three topics, effectively and clearly illustrated by specific examples presented in a highly understandable manner. Indeed the primary merit of the book resides in the author's consistent talent for presenting essential concepts in terms which should be familiar to engineers and students with only a minimal experience in automatic control theory. This attribute is achieved, not only by the manner of presentation, but also by the success in focusing only on important aspects of the theory—in other words, by an unusual selectivity of subject matter.

The phase-plane discussion focuses on construction and interpretation of the phase portrait with only a very brief and elementary discussion of optimum systems, essentially restricted to the case of second-order processes with step-function excitation. The discussion of describing functions includes two impressive and most welcome chapters on the dual input describing function. The final portion of the book presents the elementary theory underlying system characterization in terms of the response to random excitation, with the treatment similar to those by Coales and Booton.

Any strongly introductory treatment such as this involves the slighting of many topics which are considered important by many engineers—topics such as linearization, piecewise-linear analysis, the identification problem in nonlinear processes, optimum or optimalizing systems, and computer techniques for the analysis, design, and evaluation of nonlinear systems. While each of these topics is mentioned, the author has focused very strongly on the three aspects mentioned earlier. The specific objective of providing a readable, teachable, and well-motivated introduction to nonlinear control theory is exceedingly well met.

The technical production of the book seems to leave considerable to be desired, with numerous typographical errors and rather unattractive art work.

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Automatic Translation, by D. Yu. Panov

Published (1960) by Pergamon Press, Inc., 122 E. 55 St., N. Y. 22, N. Y. 65 pages+vi pages+7 appendix pages. Illus. 5⅝×8⅝. $3.50.

This small book retells some of the early work by the Russians at the Institute of Precise Mechanics and Computing Technique of the U.S.S.R., which began in 1955 and ended with work at the Institute of Scientific Information of the U.S.S.R. Academy of Sciences in 1956. Translation was started on scientific and technical material from English and other languages to Russian, using the BESM computer.

Since its publication, a number of computer techniques have appeared in the field of linguistics directed toward translation. In

the U. S., the methods of Z. Harris at the University of Pennsylvania, and of I. Rhodes at the National Bureau of Standards, have achieved especial prominence. The book in question may be considered a short and incomplete but readily comprehensible introduction to the field, for laymen. It is interesting to note that the only references given are to a well-known textbook and three articles, in English.

The most original and best part of the work is Chapter 10, "Some Scientific Questions Connected With The Problem of Automatic Translation." Here the writer points out the difference between coding and translation—changing the outward form of words, and changing the language itself. He also points out the reason why "the attempt to fit a language into some abstract scheme has proved of little effective use," because abstraction from details rejects some of the concept contained. He shows the difficulty which lies not in language structure, but in "finding equivalents for precisely those individual words. . . . " He discusses unambiguous languages, and a derived information language which would connect concepts but not form connected sentences. He concludes that the basis of concepts is an assembly of characteristics. He also concludes that the task of mechanizing bibliographical research has much in common with that of mechanizing translation, except that the former process has the peculiarity of using an artificial informational language—confined to work done within the machine—retranslated to ordinary language for the user. He points out that logical languages "trade" redundance and great length for conceptual accuracy and discusses the possibility of literary editing by machine and the use of an intermediary language for simultaneous translation into several other languages. An appendix gives an example of the BESM analysis of an English sentence for translation to Russian.

The book is worth reading for itself, and as a fragment important for its historic value. The reader is also referred to two different reviews of the work which have already appeared: by Paul Garvin and Don Swanson, in *Science*, vol. 132, pp. 343–344; August 5, 1960; and by E. G. Hill, in *J. Documentation*, vol. 16, pp. 208–211; December, 1960. These contain helpful references to further literature on machine translation.

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Council on Library Resources, Inc.
Washington, D. C.

Radio Transmitters, by Laurence F. Gray and Richard Graham

Published (1961) by McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. 427 pages+12 index pages+xi pages+20 appendix pages. Illus. 6¼×9¼. $12.50.

This book is of particular value to technicians, operating engineers and design engineers working with radio transmitters. It brings together much material that has been quite scattered in the literature. The subjects are treated in a general but clear and easy-to-understand manner with adequate depth for the intended reader. Advanced design engineers will find the references, 586 in all, at the end of each chapter of particular value if greater depth of treatment on specific topics is desired.

Transmitter requirements and characteristics for various applications and services such as telegraph, telephony, TV, radar, etc., are all covered. Frequency control techniques, RF power amplifiers and transmitter measurement techniques are given extensive treatment. The operation of conventional negative grid tubes and velocity modulated types such as klystrons and traveling-wave tubes is explained. RF coupling circuits, various types of modulation, power supplies, control and protective circuits, cooling, and RF components are covered in detail, and a short but important discussion on hazards to operating personnel is included.

Although most of the material presented is available in the literature, it has often been time-consuming to find. The authors have successfully met their objective of bringing this material together for a comprehensive treatment of radio transmitters.

WARREN B. BRUENE
Collins Radio Co.
Cedar Rapids, Iowa

RECENT BOOKS

Cotton, H., *Principles of Illumination.* John Wiley and Sons., Inc., 440 Fourth Ave., N. Y. 16, N. Y. $12.00.

Headquarters Staff of the American Radio Relay League, *The Radio Amateur's Handbook, 38th ed.* The American Radio Relay League, Inc., West Hartford 7, Conn. $3.50. Material has been revised, and a 16-page index has been included.

Kingery, W. D., *Introduction to Ceramics.* John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $15.00. General in conception and based on a rational approach to ceramic phenomena and properties; features descriptions of the microstructure and properties of ceramics as illustrative material for the application of physical ceramics to real problems.

Scandinavian Research Guide, Vols. 1–2. Scandinavian Council for Applied Research, Gaustadalleen 30, Blindern, Norway. $10.00. A directory of research institutions within technology and science, exclusive of life sciences. Gives factual information on over 1500 research institutions in Scandinavia.

Siff, Elliott J., and Emmerich, Claude, L., *An Engineering Approach to Gyroscopic Instruments.* Robert Speller and Sons, Publishers, Inc., 33 W. 42 St., N. Y. 36, N. Y. $7.50. Presents the pertinent characteristics of gyroscopes; a review of the principles and general features of existing instruments.

Truitt, T. D. and Rogers, A. E., *Basics of Analog Computers.* John F. Rider, Publisher, Inc., 116 W. 14 St., N. Y., N. Y. $12.50.

Vertregt, M., *Principles of Astronautics.* Elsvier Publishing Co., Amsterdam, Netherlands; distributed by D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. $3.50. Gives a survey of the theories underlying the different fields of astronautics, thus enabling the reader to obtain a basic knowledge of the subject.

Scanning the Transactions

Ultrasonic light modulation, discovered in 1932, has a number of interesting applications that are still not widely known. The heart of these light modulating devices is an ultrasonic cell, consisting of a liquid or solid medium which transmits both light and ultrasonic waves. A signal voltage applied to a piezoelectric transducer generates ultrasonic waves which cause periodic changes of the refractive index of the cell, thus modulating a beam of light traversing the cell. This technique was used with considerable success before the war in the development of large-screen television projectors for British cinemas. It has also been used in radar recording and display devices because it is capable of giving greater resolution than a cathode-ray tube. Ultrasonic light modulators have also been proposed for use as color modulators, frequency analyzers, signal correlators, and modulators for infrared communications systems. Perhaps we shall be hearing more of this novel technique in the near future. (A. H. Rosenthal, "Application of ultrasonic light modulation to signal recording, display, analysis and communication," IRE Trans. on Ultrasonic Engineering, March, 1961.)

Educational television has reached a point of serious consideration and considerable experimentation, judging by a number of articles that are suddently appearing. The March issue of IRE Transactions on Education alone carries three. By the time this is read, a converted DC-6 flying at 23,000 feet over Indiana will have begun experimental telecasts of high school and grade school courses to 5 million students in a 6-state area. Regular telecasts are expected to begin this fall. A detailed description, written last year, of this interesting program appears in the current issue, while a more recent interim report will appear in the 1961 IRE International Convention Record in July. The issue also describes the results of teaching a course in antenna theory to 125 electrical engineering seniors at Iowa State University and a course in electrical engineering to three sections of University of Wisconsin students, one section located at Madison and the others at Milwaukee—all by closed-circuit television.

A young man's game is an oft-heard description that can well be applied to the computer field. A survey of its membership, conducted by the PGEC, shows that the largest grouping (25 per cent) fell in the 29 to 32 age bracket, with the 33 to 36 bracket a close second (24 per cent) and 25 to 28 not far behind (17 per cent). Indeed, 84 per cent of the members were 40 or under. (K. W. Uncapher, "1960 PGEC membership report," IRE Trans. on Electronic Computers, March, 1961.)

Nonlinear circuit theory gained momentum with the pioneering contributions to Dr. Balth. van der Pol nearly forty years ago. Since then, nonlinear problems have acquired increasing importance in all branches of science. It is only fitting that a special issue of the Transactions be sponsored jointly by the Professional Group on Circuit Theory and International Scientific Radio Union and dedicated to the memory of Dr. van der Pol, a renowned scientist and educator. The issue brings together twenty-four invited articles by outstanding engineers, physicists and mathematicians from a half dozen countries. It is hoped that the publication will give impetus to the aims visualized by Dr. van der Pol serving not only as a source of inspiration for research in a field of fundamental and vital practical importance, but also as an example of international scientific cooperation. The issue contains reviews and significant original contributions in a number of areas; among them, Nonlinear Oscillations, Amplification, Synthesis, Transient Behavior, Behavior with Stochastic Signals, Automatic Control and Regulation, Stability, Differential Equations, and others. Directions of mathematical research in nonlinear circuit theory are also outlined. (IRE Trans. on Circuit Theory, December, 1960.)

Automation of Human Functions is the subject of an interesting and unusual special issue published this March by the IRE Transactions on Human Factors in Electronics. The issue gives graphic evidence of the growing importance of "intelligent machines" in the affairs of humans. The six papers in the issue speak of computer languages and of machines which aid the blind to see, take dictation, help make decisions, and solve problems. Those interested in artificial intelligence will be certain to want the cross-indexed bibliography of over 500 references on the subject which appears at the end of the issue.

A sonic Doppler navigation device the size of a car radio has been developed for providing vehicles with speed and distance measurements under circumstances where direct contact with the ground is not practical. The device is intended for use on track vehicles, such as tanks, and on helicopters, hydrofoil craft, and arctic and desert vehicles. The device is in many ways the sonic counterpart of the radar Doppler navigation systems used in aircraft. The Doppler effect, originally thought of as mainly a sonic phenomenon, has in the last four years become almost exclusively associated in our minds with electromagnetic radiation. It appears that we have now gone the full circle. (M. Wachspress, "Ultrasonic Doppler for distance measurement," IRE Trans. on Ultrasonics Engineering, March, 1961.)

Abstracts of IRE Transactions

The following issues of TRANSACTIONS have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y., at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	IRE Members	Libraries and Colleges	Non Members
Antennas and Propagation	AP-9, No. 2	$2.25	$3.25	$4.50
Audio	AU-9, No. 1	2.25	3.25	4.50
Automatic Control	AC-6, No. 1	2.25	3.25	4.50
Broadcasting	BC-7, No. 2	2.25	3.25	4.50
Education	E-4, No. 1	2.25	3.25	4.50
Electronic Computers	EC-10, No. 1	2.25	3.25	4.50
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Microwave Theory and Techniques	MTT-9, No. 1	2.25	3.25	4.50
Space Electronics and Telemetry	SET-7, No. 1	2.25	3.25	4.50
Ultrasonics Engineering	UE-8, No. 1	2.25	3.25	4.50

Antennas and Propagation

Vol. AP-9, No. 2, March, 1961

A Triangular Arrangement of Planar-Array Elements that Reduces the Number Needed—Eugene D. Sharp p. 126)

It is shown that by arranging the elements of a beam-scanning planar antenna array in a triangular pattern rather than a rectangular pattern, the number of elements needed in the array is reduced. (The number of elements needed in an array is determined from the requirement that no spurious beams form in the array pattern.) The reduction in the number of elements depends upon the solid angle over which the main beam is positioned. If the main beam is positioned within a constant angle about the array normal, then the number of elements can be reduced by 13.4 per cent by arranging the elements in a pattern of equilateral triangles rather than in a square pattern. If the main beam is positioned within a "pyramid," centered about the array normal, then the reduction is usually less than 13.4 per cent. Graphs are included showing for both element arrangements the solid angle over which the main beam can be scanned without the formation of spurious beams.

A Study of the Coma-Corrected Zoned Mirror by Diffraction Theory—S. Dasgupta and Y. T. Lo (p. 130)

A two dimensional coma-corrected zoned mirror has been investigated by a rigorous integral equation formulation, for whose solution a procedure of successive approximation can be established. For computational convenience, the exact first order solution is again approximated. The error committee in this approximation is investigated. The higher order solutions due to couplings between zones are not significant so far as the image field is concerned.

In the example studied, it is found that for a system with small F-number and an illumination of small taper the coma aberration of this zoned mirror is practically eliminated for scan angles up to 25°, in sharp contrast to a smooth parabola whose scanning characteristics are also presented for comparison. However, the chromatic aberration overshadows this advantage for a wide-band application; the total bandwidth of this example is only a few per cent.

Microwave Antennas Derived from the Cassegrain Telescopes—Peter W. Hannan (p. 140)

A microwave antenna can be designed in the form of two reflecting dishes and a feed, based on the principle of the Cassegrain optical telescope. There are a variety of shapes and sizes available, all described by the same set of equations. The essential performance of a Cassegrain double-reflector system may be easily analyzed by means of the equivalent-parabola single-reflector concept.

Techniques are available for reducing the aperture blocking by the sub dish of the Cassegrain system: one method minimizes the blocking by optimizing the geometry of the feed and sub dish; other methods avoid the blocking by means of polarization-twisting schemes. The former method yields good performance in a simple Cassegrain antenna when the beamwidth is about 1° or less. The latter methods are available for any application not requiring polarization diversity, and an optimized set of polarization-operative surfaces has been developed for these twisting Cassegrain antennas.

Experimental results, presented for practical antennas of both types, illustrate the feasibility of these principles. A number of unusual benefits have been obtained in the various Cassegrain antenna designs, and additional interesting features remain to be exploited.

Multiple Beams from Linear Arrays—J. P. Shelton and K. S. Kelleher (p. 154)

The problem of devising a passive RF transmission line feed system to provide independent multiple outputs from a linear array is considered. It is shown that a lossless, matched feed network is possible only for uniform aperture distribution. The general feed system for connecting 2^n inputs to 2^n elements is shown to consist of conventional hybrid junctions with associated phase shifters. In order to increase the possible number of elements in the array, the problem of finding applicable junctions more complex than the hybrid is considered. Junctions with three inputs and three outputs and with four inputs and four outputs are derived for use in multiple feed networks, expanding the number of elements in the array to $2^l\,3^m\,4^n$.

A New Technique for Electronic Scanning—H. E. Shanks (p. 162)

The concept of time-modulated antennas has recently been demonstrated as a means of overcoming many of the limitations currently restricting advances in the antenna art. Of special importance is the mathematical possibility of generating a pattern complex capable of providing simultaneous scan operation. This characteristic is realized by periodic time modulation of the aperture distribution. This paper discusses the theory of simultaneous "scanning" using time modulation techniques and shows that the required pattern complex is generated by a progressive-pulse aperture excitation. The fundamental equations and relationships concerning the form of pulse excitation and "scanning" coverage are derived . In addition, practical methods of physically generating the proper pulse-excited aperture are described, and the necessary detection requirements are delineated.

Cylindrical Shields—R. W. P. King and C. W. Harrison, Jr. (p. 166)

The effectiveness of an imperfectly-conducting cylindrical shield of small cross section depends on both the attenuation through the metal wall of the externally maintained field and the amplitude of the current that is induced in the cylinder. When the length of the cylinder, which behaves like a linear scattering antenna, approaches a resonant value, the currents induced in the walls and the field inside the tube are relatively large. Under these conditions, large currents may be induced in a thin dipole placed coaxially within the shield.

Folded Dipoles and Loops—C. W. Harrison, Jr. and R. W. P. King (p. 171)

In Section I the theory of linear arrays consisting of two or more closely spaced elements that are interconnected by lumped reactances is reviewed. Specific application is made to two-element end-loaded folded dipoles and monopoles constructed of conductors with different diameters, to series tuned three-wire folded dipoles and monopoles, and to a three-wire-line reactor and impedance transformer. In Section II the circular folded dipole or Halo antenna is treated.

Sidelobe Reaction by Nonuniform Element Spacing—Roger F. Harrington (p. 187)

A perturbational procedure for reducing the sidelobe level of discrete linear arrays with uniform amplitude excitation by using nonuniform element spacing is presented. The calculation of the required element spacings is quite simple. The method can reduce the sidelobe level to about $2/N$ times the field intensity of the main lobe, where N is the total number of elements, without increasing the beamwidth of the main lobe. Several examples are given.

Photoconductive Modulation of Microwave Electric Fields—W. E. Bulman, B. C. Potts, and R. B. Green (p. 193)

Thin films and coatings of photoconductive cadmium sulfide material are being used to modulate electric fields. This paper analyzes these films in terms of the physics involved and

the effect upon microwave fields. Experimental procedures show the validity of the analyses and indicate possible applications of the principle. Practical significance of the technique is illustrated by experimentally determining the areas of a particular flat plate that produce the first sidelobe of the scattering pattern.

Diffraction of a Plane Wave by an Infinite Slit in a Unidirectionally Conducting Screen—S. R. Seshadri (p. 199)

The scattering of a plane electromagnetic wave of wave number k, by an infinite slit (of width $2a$) formed by two unidirectionally conducting, semi-infinite coplanar screens of zero thickness is considered. By employing an integral equation procedure, a rigorous asymptotic solution is obtained up to order $(ka)^{-5/2}$. The currents induced on the screens and the first few terms in the transmission coefficient are evaluated. The similarity between this and the corresponding problem involving perfectly conducting screens is pointed out.

Refraction Compensation in a Spherically Stratified Ionosphere—Stephen M. Harris (p. 207)

For satellite orbit determinations, it is necessary to know the range to the satellite with great accuracy. The presence of the ionosphere with a frequency-dependent index of refraction produces errors in range measurements made by either CW (Doppler) or pulse radio techniques. A measurement scheme is proposed that gives the instantaneously-corrected range for a spherically stratified ionosphere without recourse to any further assumptions about the electron density profile. The corrected range is given by the average of the CW and pulse range measurements and is free of the first-order error contributed by the inverse frequency-squared term in the refractive index. This scheme is shown to be slightly more effective than a scheme combining the results of two CW range measurements. Expressions are also derived for the higher-order residual errors which remain after the proposed compensation, and the expressions for the variation in arrival angle with frequency are given. These expressions depend upon the integrated effect of the free electrons; consequently, a calculation with a simple profile should yield typical results. The compensated range error and the residual errors are given for a satellite at a height of 640 km, arrival angles in the first quadrant, and at a frequency of 100 Mc This compensation scheme cancels all but a few per cent of the original range error.

Scatter Communications with Radar Chaff—R. A. Hessemer, Jr. (p. 211)

The first part of this paper is concerned with finding an analytical expression for the scattering cross section of chaff oriented randomly within a vertical cone. The dipoles are allowed to take on all the angles within this cone. A vertically-polarized receiver is assumed off on the horizon and the transmitter on the ground below the chaff. The cross section is a function of the conical angle of the configuration and the angle between a normal to the ground and the incident electric field from the transmitter. Fig. 2 is a plot of the scattering cross section as a function of these two angles.

Half-wave chaff randomly distributed within a conical angle about a vertical is not the most effective ensemble, but is a practical one at the lower frequencies. Cutting all these half-wave dipoles into very short ones makes it practical to place them in a horizontal position which has an ensemble gain over the conical but a reradiation loss, since short dipoles are less effective scatterers than half-wave ones. The second part of this paper compares the reradiation loss and horizontal ensemble gain.

Diffraction by a Slit—Robert Plonsey (p. 217)

The electric field diffracted by a strip caused by an incident cylindrical wave with E parallel to the edge, at various angles of incidence, is measured in a parallel plane medium. The field is compared with that computed from geometrical optics currents and with the addition of equivalent line currents at the edges. The edge "line currents" improve the geometrical optics current field particularly at oblique incidence.

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Audio

Vol. AU-9, No. 1, January–February, 1961

The Editor's Corner—Marvin Camras (p. 1)
PGA News (p. 2)

Hall Effect Wattmeters—D. P. Kanellakos, R. P. Schuck, and A. C. Todd (p. 5)

A transmission-type wattmerer can be formed utilizing the Hall effect to give wide-range multiplication over a frequency range extending from the low audio-frequency region to the SHF band. At the low frequencies, the Hall device can take the form of a thin wafer of indium antimonide mounted in series with the center leg of a cup core magnetic circuit. An audio-frequency wattmeter employing such a structure has been analyzed, constructed and tested. The wattmeter has a range of zero to 400 milliwatts when used in a 600-ohm circuit, and displays a frequency error of less than plus and minus 3.5 per cent over a frequency range of 100 to 6000 cps for a unity power factor load.

In the SHF range, a cavity and electric probe are arranged to provide excitation for the indium antimonide wafer in the manner first employed by Barlow. The cavity, about one wavelength long, is separated from the waveguide in which power flow is to be measured by a thin brass wall, and is coupled to the waveguide by a rectangular slot in the top wall of the waveguide. The semiconductor wafer is placed in the center of the cavity, at a position having a maximum in magnetic field and a minimum in electric field for a given cavity excitation value. The electric probe, which extends from the cavity to the waveguide, provides excitation current for the semiconductor element. Three tuning adjustments are required, two for the cavity and one for the probe. A wattmeter embodying this technique has been found to provide a sensitivity of one dc microvolt per milliwatt of transmitted power at a frequency of 9.40 Gc.

The Effects of Track Width in Magnetic Recording—D. F. Eldridge and A. Baaba (p. 10)

The effects of track width on various performance characteristics have been measured over a wide range of widths. Signal level, noise, and signal-to-noise ratio were determined for track widths from 1.1 mils to 92 mils. The effects of crosstalk, actual recorded track width vs head width, and tape guiding, are described. The experimental data are in good agreement with theory, and no serious practical limitations on the use of very narrow tracks were discovered. High-density audio and pulse recordings were made without difficulty. Digital bit densities of one million per square inch and above are shown to be possible.

Theory of Motional Feedback—Egbert de Boer (p. 15)

In motional feedback the *mechanical vibrations* of the loudspeaker cone are the source of the feedback voltage. Feedback then improves the over-all response characteristic and reduces the total distortion. The theory of this method is presented here in a simplified, though enlightening, way. The treatment is based on an unorthodox theorem on impedance conversion by feedback.

A Stereophonic Transistor Preamplifier—Werner Steiger (p. 22)

Designed to drive two transistor-power amplifiers previously described, this preamplifier contains all the controls necessary for stereophonic and monophonic programs. The first two of the three stages form a complementary circuit readily adaptable for the various input sources, which provides the necessary gain and equalization for low-level phonograph and tape inputs, guard voltages to reduce the capacitive effects of the input cables, and impedance transformation for high-level inputs from tuners and other tube circuitry. Loudness control is achieved by the combined effect of variable negative feedback and a novel tone-control arrangement. The frequency range, including the power amplifiers, extends from below 6 cycles to above 50 kilocycles per second.

Polarity, Phase and Geometry—Paul W. Klipsch (p. 25)

Following the teaching of the early Bell Telephone Laboratories' experiments in auditory perspective, wide speaker spacing is needed to realize accuracy of geometry. This results in substantially random phase, so that polarity is relatively unimportant. This is supported by Lissajous figures of two-channel stereo signals.

It is still good practice, however, to observe polarities, if for no other reason than to permit monophonic reproduction over a stereophonic array. Where the stereo signals contain a strong monophonic component, correct system polarity is better than random polarity. In some stereo situations, bass is improved by correct polarities.

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Automatic Control

Vol. AC-6, No. 1, February, 1961

Message from Moscow—The Editor (p. 1)
Chairman's Report—J. M. Salzer (p. 3)
The Issue in Brief (p. 4)

An Optimal Strategy for a Saturating Sampled-Data System—C. A. Desoer and J. Wing (p. 5)

Consider the usual sampled-data control system in which the sampler is followed by a zero-order hold and the transfer function is $G(s)=1/s(s+a)$. Saturation is represented by the fact that the forcing function applied to $G(s)$ may not be larger than 1 in absolute value. The problem is to determine a saturating zero-order hold forcing function which forces the system from an arbitrary initial state to equilibrium in the least number of sampling periods. Such a forcing function is defined as an optimal strategy.

The state plane is divided into boundary states and interior states. To each boundary state corresponds a unique optimal strategy. To each interior state correspond infinitely many optimal strategies.

From the system parameters a polygonal curve, called the critical curve is defined in the state plane. An optimal strategy is then proposed in which the required forcing function is simply obtained by computing the distance of the representative point in state plane to the critical curve. A simple computer is proposed to implement this optimal strategy. Finally, the proposed optimal strategy is shown to reduce in the limit as $T\rightarrow 0$ to that of the corresponding continuous system.

Time-Optimal Control of Higher-Order Systems—Fred B. Smith, Jr. (p. 16)

Practical extension of time-optimal control to systems of higher order than three has been limited primarily by difficulties in physically representing surfaces in a phase space of these higher dimensions. A method is presented here for obtaining the forcing function as a function of the state variables without requiring use of the phase space concept. On line solution of a

set of transcendental equations is required. Results of a digital simulation of a fourth-order, real-root, single-degree-of-freedom system are presented. In a digital solution the system operates as a series of short open-loop control intervals. The effect of including derivatives of the input for prediction is shown for second-order model inputs.

Control System Performance Measures: Past, Present, and Future—W. C. Schultz and V. C. Rideout (p. 22)

An increased amount of emphasis on the mathematical formulation of control system performance can be found in recent literature on automatic control. There are two areas of control system theory in which the application of performance measures is of interest: 1) the evaluation of control system designs in general and 2) the design of adaptive control systems. In the former case, the performance measure is becoming an increasingly important aid to the control system designer. In the latter case, the performance measure takes on even greater significance, since adaptive systems include, by definition, a performance measure as an essential function which permits correction of system dynamic response during actual operation. Furthermore, the over-all evaluation of the adaptive loop itself presents new problems in the choice and use of performance criteria.

In the past, emphasis has been placed on various types of integrals, such as integral of error-squared and integral of the product of time and absolute error (ITAE); present emphasis is being placed on forms of integrals of a more general type; the trend for future emphasis appears to be in applications of statistical concepts and in attacking the problem of choice of the error measure in the adaptive system.

A Simulator Study of a Two-Parameter Adaptive System—R. J. McGrath and V. C. Rideout (p. 35)

The use of sinusoidal parameter perturbation applied to a feedback control system to obtain an adaptive scheme which optimizes the system for changes in inputs and/or system parameters is discussed. It is shown that if a parameter perturbation signal is cross-correlated with the system error squared, the correlator output can be used to adjust the parameter to minimize the mean-square error. Other error measures may also be used. Two or more parameters may be simultaneously adjusted if they are perturbed at different frequencies, and each provided with an independent adaptive loop. A computer simulation of a third-order system having two adjustable parameters was examined for a variety of inputs including random signals. It is shown that the scheme minimizes the mean-square error in all cases.

Optimum Prediction with a Mean Weighted Square Error Criterion—Clarence C. Glover (p. 43)

The linear prediction theory is examined using a mean weighted square error criterion. A specific nondeterministic weighting function is used. The problem is reduced to that of solving integral equations which are written in terms of correlation functions which can be calculated by averaging over the ensemble. A complete solution is given for the problem using Gaussian statistics with no correlation between noise and true signal.

Control Systems with Minimum Spectral Bandwidth of Plant Input—James C. Hung (p. 49)

A design method for control systems that minimizes the spectral bandwidth of the plant input signal is discussed. The plant input signal is minimized subject to the constraint that the integral square error for deterministic inputs or the mean square error for random inputs be limited to a known desired value.

The control system transfer function that satisfies these requirements is derived, and the functions used in bandwidth shaping are discussed. An example of a system design using this technique is given.

A Network Theory for Carrier-Suppressed Modulated Systems—Gerald Weiss (p. 54)

The performance of linear networks in the presence of carrier-suppressed modulation is re-examined in the light of the latest advances in theory. Both analysis and synthesis methods are presented.

Design Aspects of Attitude Control Systems—M. F. Marx (p. 67)

Figures of merit, besides those of performance, are discussed relative to the attitude control system of a vehicle capable of leaving and returning to the atmosphere. In addition to extreme changes in flight condition, these applications are subject to variations in configuration and performance requirements.

Traditionally, control optimization has been concerned with minimizing or maximizing a variable system function such as error. Quite often these error criteria are replaced by other criteria such as invariance and the capacity for adaptability. In fact, during a complete mission including exit to re-entry it is desirable to utilize variable figures of merit.

Examination of the control requirements of a modern returnable space vehicle illustrates how the various figures of merit dictate the design configuration. In those phases of the mission where self-adaptive control is employed, the figure of merit is usually determined by the particular technique selected. It is further demonstrated how the figure of merit varies with the mission phase as the control actuation configuration changes.

Analysis and Design of Feedback Systems with Gain and Time Constant Variations—Kan Chen (p. 73)

The design of a feedback control system containing an element with proportional variation of gain and time constant is a common problem encountered by control engineers in practice. The problem includes the stabilization of a system, which is open-loop unstable when both the gain and the time constant of the element are negative. This paper presents a method for analyzing the transient response of systems containing the aforementioned element, and designing the systems to meet transient specifications.

Evaluation of Transient Response Coefficients—D. S. Billingsley and M. G. Rekoff, Jr. (p. 80)

A method is presented for evaluating the transient response coefficients of response functions having poles of any multiplicity. The calculations are effected by use of a recursion-type formula with data obtained from distance and angle measurements on the root-locus plot of the system being analyzed.

Correction to "On Optimal and Suboptimal Policies in the Choice Control Forces for Final Value Systems"—Masanao Aoki (p. 83)

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Information on Translations of Russian Technical Journals (p. 91)

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Broadcasting

Vol. BC-7, No. 2, March, 1961

Scanning This Issue—The Editor (p. 1)

FCC Rule Making—J. G. Rountree (p. 2)

Use of Digital Computer to Determine Broadcast Station Nighttime Interference—Israel Akerman (p. 5)

A computer program developed and used by the Department of Transport to determine nighttime sky wave signal interference to ground wave service is described. The formulas and sequences used by the program are discussed. A sample calculation of the program is given. The advantages of such a program, to the broadcast design engineer, are accuracy, reliability, speed and convenience.

Experiments with a Slope-Feedback Coder for Television Compression—M. P. Beddoes (p. 12)

Experiments with a slope-feedback variable-velocity-scanning method for compressing the bandwidth of television signals are described. The test conditions were intended to demonstrate the method under favorable conditions and one test-pattern of moderate complexity—test-card "C"—only was used. The experiments showed that smoothing of the signal by narrow bandwidth can be overcome using the method. But, even at best (corresponding to the least compression ratio observed = 1.7/1), spatial distortion of the received picture is noticeable.

The reason for the spatial distortion seems, by a process of elimination, not to be due to the effects of noise but to be characteristic of the method itself; it would occur even in a theoretically noiseless system because the "slope-signal" controlling the spot movement at the transmitter differs from the corresponding signal at the receiver which has added to it anticipatory and overshoot ripples in the vision channel. This is a basic fault and rules out the application of slope-feedback compression to pictures of the complexity of test-card "C."

Digital Computers for Television Automatic Switching Control—A. B. Ettlinger (p. 29)

The first application, so far as is known, of a digital computer to television automatic switching control is presently in development under CBS auspices. The resulting system offers a considerable advance over previously-used techniques with respect to functional capacity, ease of revision of program schedule, and adaptability to changes in operating practice. In meeting these stringent requirements the computer approach proves to be economically sound, offering features unattainable by other means except at prohibitive cost.

Education

Vol. E-4, No. 1, March, 1961

Editorial—Wilbur R. LePage (p. 1)

The Paradox of Education in Industry—Yates M. Hill (p. 3)

To some, the fact of engineering education in industry may represent a paradox. It has been introduced in order to make it practical for the engineer to find a satisfying division of effort between pursuing immediate job goals and longer-term knowledge objectives. When one considers the economic objectives of industry, the objectives and responsibilities of engineers, and the effects of scientific and technological progress, the continued broad education of engineers seems a necessity.

The discussion is intended to be helpful to individual engineers in determining their own knowledge objectives in connection with "engineering" their own personal development.

Thoughts on Engineering Education—Norman Balabanian (p. 6)

Engineering curricula are in a process of dynamic change as a result of the rapidly accelerating rate of advance technology. The rapid accumulation of knowledge leads to a selection process of the materials to be included in the curriculum. If the emphasis in engineering education is on the imparting of facts—even facts about "fundamentals"—the engineer will soon be wallowing in the wake of the advancing technology. Major emphasis in engineering education should be placed on developing in the student the open-minded attitudes and the

critical approach to the solving of problems—be they technical, economic, political, or social problems—that is implicit in the phrase "method of science."

A New Approach to Electrical Engineering Laboratories—Sheldon Plotkin (p. 9)

A short analysis of the purposes and achievements of electrical engineering laboratories is given. As a result of this analysis, the conclusion is drawn that laboratories, as presently conceived, fail to accomplish their objectives. The new approach proposed is to limit the objectives as well as the laboratory time, which would result in an advanced electrical measurements course with an associated laboratory to demonstrate the lecture material.

A Course in Engineering Measurements—Peter K. Stein (p. 12)

This paper describes the first course which engineering students at Arizona State University take in the field of measurements. This Senior course is part of the "core," *i.e.*, it is taken by all engineering seniors (except electrical engineers who may take it as an elective subject).

A Controlled-Temperature Device for Transistor Tests—E. F. King and F. L. Walker (p. 21)

A laboratory device for operating transistors at selected, known temperatures in the 15–95 degrees C range is described. After a simple, one-time calibration no further temperature measurements are necessary. The unit is simple, rugged and easily constructed.

The Role of an AC Network Calculator in the Electrical Engineering Curriculum at The Pennsylvania State University—P. E. Shields (p. 22)

Changes are being made in the content of present-day electrical engineering curriculums which require a fresh, modern outlook with regard to techniques pertaining to problem assignments and laboratory projects. Course content is being upgraded and broadened in the direction of a fundamental analytical approach. The recent revision of the electrical engineering curriculum at The Pennsylvania State University, University Park, includes liberal use of an ac network calculator in all electrical engineering options, which are outlined in this paper. The uniqueness of the ac network calculator at the University is that it is used exclusively for educational purposes, especially in the Department of Electrical Engineering. This paper discusses the "when, where, and how" of the utilization of this facility in the electrical engineering curriculum.

Application of Airborne Television to Public Education—C. E. Nobles, F. G. Mullins, Jr., J. L. Wagner, and R. Lee (p. 26)

In an effort to bring the advantages of television teaching to the maximum number of public schools, particularly those in hilly and sparsely settled parts of the country, a project has been launched to provide airborne transmission of six simultaneous school lesson programs. Coverage area is increased approximately 20 to 1 by this means compared to the usual ground station coverage. Airborne transmission is the most economical and effective use of the available (UHF) frequency spectrum.

Results already established regarding the use of television as a teaching medium are projected via an airplane flying at 23,000 feet altitude over an area covering parts of six Midwestern States. Educational advantages and problems inherent in ground station TV are intensified in this larger coverage area, and planning is governed accordingly. Technical problems include multiplex transmission of six lessons simultaneously, keeping the airplane on station, accommodation of transmitting equipment in the airplane, and determination of bandwidth for optimum picture quality.

Teaching an Advanced Electrical Engineering Course by Closed-Circuit Television—W. L. Hughes and C. J. Triska (p. 33)

A course in antenna theory was taught to about 125 electrical engineering seniors through the use of closed-circuit television. The techniques of instruction followed were a composite of those used in other television instruction experiments, and the details are given in the paper. Nonedited student comments and an evaluation of the results are given. Closed-circuit television may well have some fruitful application for this level of instruction.

A Description of a Simple Teaching Machine—B. James Ley (p. 38)

The operation of the electrical circuit of a relatively simple teaching machine and some ways in which this machine has been used are described.

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Electronic Computers

Vol. EC-10, No. 1, March, 1961

Unate Truth Functions—Robert McNaughton (p. 1)

This paper contains some applications of an elementary study of unate truth functions. One application is a method of deciding when a truth function is linearly separated, *i.e.*, is expressible as a linear polynomial inequality in its arguments (letting 1 represent truth and 0 represent falsity). Other applications are to contact nets and to rectifier nets. Much of the material of this paper, although not in print, is well known to some logicians and switching theorists. Nothing from the first three sections is original.

Linear-Input Logic—Robert C. Minnick (p. 6)

Techniques are developed for the logical design of magnetic core circuits to produce arbitrary single-output combinational switching functions. The approach is based on the relationship of a single magnetic core circuit to a linearly separable switching function. A synthesis procedure is developed which uses a pair of logical primitives, AND with NOT and OR with NOT, which are similar to the STROKE primitive and its inverse. Procedures are developed for the synthesis of symmetric functions which require no more than the integral part of $(n+3)/2$ cores, approximately half the number used in previously published procedures. The synthesis of arbitrary switching circuits is treated as a linear programming problem, and a table of all four-variable circuits is presented in which no circuit requires more than three cores.

Axiomatic Majority-Decision Logic—M. Cohn and R. Lindaman (p. 17)

An algebra suited to logical design with majority-decision elements (parametrons, Esaki diodes, etc.) is developed axiomatically. The utility of the new algebra is demonstrated by resolving sample problems.

Computer Design of Multiple-Output Logical Networks—Thomas C. Bartee (p. 21)

An important step in the design of digital machines lies in the derivation of the Boolean expressions which describe the combinational logical networks in the system. Emphasis is generally placed upon deriving expressions which are minimal according to tome criteria. A computer program has been prepared which automatically derives a set of minimal Boolean expressions describing a given logical network with multiple-output lines. The program accepts punched cards listing the in-out relations for the network, and then prints a list of expressions which are minimal according to a selected one of three criteria. This paper describes the basic design procedure and the criterial for minimality.

Games That Teach the Fundamentals of Computer Operation—Douglas C. Engelbart (p. 31)

One who wishes to give a group of laymen a feeling for the way we computer engineers can coax sophisticated information-handling behavior from an organization of simple physical elements can provide a striking on-the-spot example by training his laymen to simulate various kinds of simple elements and by organizing them into a network whose behavior is obviously more sophisticated than that of any element. Each individual watches the up-down hand position of one or two others, and adjusts his own hand position according to a response task which is equivalent to that of an AND, OR, NOT, or flip-flop element—although task assignments are made in such a way that the participants don't hear a single esoteric word, nor realize that they might be doing "logic." Counters, shift registers, and adders may be organized and operated in a way which proves very entertaining to participants and onlookers, and yet which provides them with very realistic basic concepts about how a computer might work.

Bilateral Switching Using Nonsymmetric Elements—M. Aoki and G. Estrin (p. 42)

Magnetic-core memory elements characteristically require bipolar applied fields. The vanishing inner diameter of toroids and the loss of the third dimension entirely in deposited thin films demands minimization of the number of wires.

A configuration which has been investigated and applied in a word organized memory at the University of California at Los Angeles is illustrated in Fig. 2. It consists of a pair of mutually inverted and parallel connected transistors. The transistors are not in general symmetrical.

This paper discusses some of the system considerations which determine the important design parameters. Methods for location of regions of satisfactory operation in the many-variable space of the inverted transistor pair are described.

Although a particular design problem is discussed, attention is focused on the question, "What classical and new procedures can we use to reduce the number of dimensions in such design problems?" The power of the computer as a design tool is crucially dependent upon such processes.

Ferrite Toroid Core Circuit Analysis—R. Betts and G. Bishop (p. 51)

An analysis of the terminal characteristics of thin ferrite toroid cores under arbitrary drive and load conditions is presented. The analysis is founded only on the following two experimentally confirmed conditions: 1) the time required for a complete reversal of flux under unloaded conditions is inversely proportional to the magnitude of a step-driving field which is in excess of the critical field required to initiate flux change; 2) the open circuit voltage-time output waveforms caused by step driving currents are identical when normalized with respect to amplitude and time.

The normalized output voltage waveform $f'(x)$ is used to develop a terminal characteristic equation. It is shown that $f'(x)$ may be obtained by using a nonideal step-input current. Utilizing a modified Gaussian equation to represent $f'(x)$, equations are developed to allow the prediction of core response to arbitrary input waveforms, using 4 parameters easily obtained from voltage response vs NI step-drive plots, and $f(x)$, which is the integral of the normalized expression for the open circuit voltage $f'(x)$, and is proportional to the flux switched in the core.

The equations are expanded to include a load circuit and to test the validity of the expressions developed. Theoretical and experimental results are compared for a core loaded with series RL and RLC circuits with both ramp and step-drive currents. Agreement is shown to be good, even though the core used was not particularly thin.

A Digital Static Magnetic Wire Storage with Nondestructive Read-Out—C. G. Shook (p. 56)

After a brief review of pertinent magnetic effects and sonic wave propagation in elastic media, a nonvolatile, digital, magnetic storage scheme is described, wherein binary words may be stored by magnetizing segments of a wire, and the information may be read out an unlimited number of times with no deterioration of the stored information. Two storage schemes are presented: a temporary, electrically addressed storage, and a permanent, program-type store. Bit-storage density, read-out and input pulse shapes, and read-out frequency are noted. Possible limitations such as losses, temperature effects, and pulse shape are balanced against advantages and a comparison is made to a number of other types of bit storage.

Correction to "Minimization of Contact Networks Subject to Reliability Specifications"—Arthur Gill (p. 62)

A Digital Correlator Based on the Residue Number Systems—Philip W. Cheney (p. 63)

A system design for a digital correlator based on the application of the residue number system for computation is presented. Areas of investigation include sampling, analog-to-residue conversion, logical design of the arithmetic units, residue-to-analog conversion, and modes of operation of the proposed digital correlator. The advantages of speed of computation and simplicity of logic due to the use of a residue number system are shown to result in a significantly faster and simpler system than if a conventional number system were used. The resulting digital correlator is designed for megacycle sampling and computation with a 0.1 per cent system precision.

A Function Generator Using Cold-Cathode Selector Tubes—R. M. Duffy and C. P. Gilbert (p. 71)

A method of generating voltages which are arbitrary functions of time is suggested in which a chain of cold-cathode selector tubes is used as a single-pole, multi-position switch: accuracies of ± 1 per cent can be achieved with relatively simple adjustment.

A generator using this method is described in detail, and typical output curves are shown. The generator is extremely versatile, not only due to the range of functions which can be produced, but also due to its ability to:

1) operate over a wide range of speeds,
2) change instantaneously from one speed to another, and
3) generate two separate functions, one displaced with respect to the other by a variable, preset delay.

Initial Conditions in Computer Simulation—K. S. Miller and J. B. Walsh (p. 78)

A technique is developed for the straightforward simulation of the transfer function of a certain class of linear systems. This method is particularly well adapted to the analysis of systems with fixed transfer function and variable initial conditions and forcing functions. In particular, a single simulation, minimal in its use of integrators, will suffice to handle forcing functions and initial conditions on both input and output.

1960 PGEC Membership Report—Keith W. Uncapher (p. 81)

The third biennial PGEC Membership Survey was completed in the Fall of 1960. The response to the questionnaire was excellent (58 per cent), and the data extracted are reported herein. Factors regarding the nature of work, geographic location, salary, education, and fringe benefits are reported.

Bar graphs, curves, and commentary are in cluded to aid in assessing the Professional Group's character and growth.

Correspondence (p. 92)
Contributors (p. 101)
Reviews of Books and Papers in the Computer Field (p. 105)
Abstracts of Current Computer Literature (p. 118)
PGEC News (p. 142)
Notices (p. 143)

Human Factors in Electronics

Vol. HFE-2, No. 1, March, 1961

Progress in Artificial Intelligence—Thomas Marill (p. 2)

Computer Languages for Symbol Manipulation—Bert F. Green, Jr. (p. 3)

Complex, flexible, computer programs can be written easily in list-processing languages. Storage registers are linked together in arbitrary sequences to form lists and list structures, which are the units of the languages. Special provisions are made for recursive subroutines and for hierarchical programs. These particular languages have been used to write game-playing, problem-solving, and other "intelligent" programs.

Reading Machines for the Blind—H. Freiberger and E. F. Murphy (p. 8)

Methods of reading for the blind are discussed, mindful of the human-factors and systems-engineering concepts involved. Machines to allow independent reading of ink-print by the blind are classified in terms of cost, complexity, auditory or tactile output, usefulness, ease of operation, and learning problems. Current experiments in character recognition are briefly mentioned, and in conclusion reference is made to using the technical advances on related problems to speed developments in this field. Also mentioned are the widespread application of the benefits of current research to both civilian and veteran populations and the probable need for prescription teams when devices become available for routine use.

Man-Computer Cooperation in Decisions Requiring Common Sense—D. B. Yntema and W. S. Torgerson (p. 20)

Men and Computers could cooperate more efficiently in real-time systems—and perhaps in long-range planning too—if a man could tell the Computers how he wanted decisions made, and then let the machine make the decisions for him. In the next few years there will probably be considerable pressure on system designers to adopt such arrangements. The problem of enabling a man to convey his decision rules to a machine will in many cases prove less formidable than it might at first appear. Three methods are discussed. As experience with man-machine cooperation of this type accumulates, problems for research will be generated. An attempt is made to foresee what some of them will be.

Programming Intelligent Problem Solvers—Walter R. Reitman (p. 26)

How do problem-solving programs work? How much has been accomplished with them so far? What kinds of developments may be anticipated over the next few years? Two research programs under development are discussed as illustrations of the evolution of heuristic programming systems. Applications of these techniques in studies of problem solving in mathematics and symbolic logic, in industrial and business problems, in laboratory tasks, chess playing and the understanding of language also are considered. The emphasis throughout is on exposition of methods and goals. The paper concludes with a discussion of longer range problems, aspirations, and research strategies in work on artificial intelligence.

Automatic Recognition of Speech—Thomas Marill (p. 34)

Research in the field of automatic speech recognition is reviewed. Despite the considerable effort which has been devoted to this field, the results are still quite limited.

The devices reviewed have achieved only very small recognition vocabularies (the ten spoken digits, for example). Those devices which do not make use of a computer are further restricted to the recognition of the speech of a particular speaker, for whose voice they have been specially adjusted. This limitation is being overcome by the more complex equipment which incorporates a digital computer.

A Selected Descriptor-Indexed Bibliography to the Literature on Artificial Intelligence—Marvin Minsky (p. 39)

Communications (p. 56)
Reviews of Current Literature (p. 61)
Contributors (p. 64)

Industrial Electronics

Vol. IE-7, No. 3, December, 1960

Applications of Transfluxors to an Electromechanical Control System—Paul Pargas (p. 1)

A control system using multiaperture magnetic cores of the transfluxor type provides a reliable method of commanding paper advance and auxiliary functions in a photographic printer. The transfluxors are operated in a mode which permits larger variations in drive control pulses and which is less sensitive to extraneous signals. Coupling to the mechanical system is done via thyratrons to clutches and relays. Advantages over a purely relay or flip-flop logic for this control purpose are discussed briefly.

Self-Triggering Spark Gap "Crowbar"—Victor Wouk (p. 7)

A self-triggering spark gap has been developed for use as a simple "crowbar" at the output of a high-voltage dc power supply to protect the load in case of load failure.

The system consists of three spheres arranged in a triggered spark gap system, with the center sphere maintained at a fixed potential with respect to the other two spheres (one at ground, one at high voltage) under steady-state conditions. When the load fails, capacitive coupling drives the triggering sphere to a voltage that causes breakdown between the high-voltage sphere and the triggering sphere, with subsequent firing of the entire protective gap system.

Advantages over thyratrons for this "crowbar" application include the following:

1) the voltage rating is unlimited,
2) the current rating is of the order of tens of thousands of amperes,
3) reliability is greater than that of a thyratron, due to lack of filament supply requirements, auxiliary triggering circuits, etc.,
4) less costly than the thyratron circuit.

Problems Encountered and Solved in Starting Up of Computer Control Systems—Robert P. Adams (p. 10)

Techniques for improving the accuracy and safety of computer-controlled systems are presented. The scope of the paper is limited to those techniques which are being used successfully in operating computer-controlled processes.

Dual-Purpose Synchronizing and Dependent Drive Time-Sharing for Point-to-Point Numerical Control—George W. Younkin (p. 15)

Machine tool drives, used in conjunction with numerical positioning, can generally be

classified into two categories; some form of servo drive or an open-loop drive such as a multiple gear train and on-off clutches. Dual-purpose synchronization will permit the use of either form of machine drive with the same numerical control equipment. By time-sharing one clutch drive system between two axes of a machine, it is possible to approach the performance of a separate machine drive for each axis at the cost of a a single drive.

A Precision Gamma-Ray Absorption Meter—Phillip L. Jessen (p. 21)

A system which measures the absorption of the gamma radiations of Cobalt-60 in various materials is described. A thickness measurement device has been included in the system so that density (or specific gravity) of the sample materials may be automatically calculated. Accuracies of ± 0.03 per cent for the specific gravity measurement are obtained.

Microwave Theory and Techniques

Vol. MTT-9, No. 1, January, 1961

Frontispiece, 1959 Microwave Prize (p. 2)

UHF Strip Transmission Line Hybrid Junction—I. Tatsuguchi (p. 3)

A hybrid junction has been developed using a symmetrical strip transmission line for application in the UHF range. It has a frequency band of ± 20 per cent where the input voltage standing-wave ratios at all ports are less than 1.26 (2 db), the power divisions are within 0.1 db, and the difference in power between the series input and parallel input ports is less than 0.3 db. The isolation is greater than 40 db and 24 db, respectively, for the two pairs of conjugate ports. These circuits are relatively small, light-weight, simple to build and reproduce, and are inexpensive.

The approximate equivalent circuit of the configuration assuming transmission in the TEM mode is presented. The results of the analysis and the important features in the design and fabrication and a few modifications of the configuration are discussed.

Gallium-Arsenide Point-Contact Diodes—W. M. Sharpless (p. 6)

This paper describes some of the work on gallium-arsenide point-contact diodes which is currently in progress at the Bell Telephone Laboratories, Holmdel, N. J. Gallium arsenide, one of the Group III-V intermetallic compounds, possesses properties which tend to make it superior to either silicon or germanium for many high-frequency diode applications. By controlling the resistivity of the gallium arsenide and the point-contact processing techniques, diodes have been fabricated specifically for use as millimeter wave first detectors, high-speed switches, and reactive elements for microwave parametric oscillators and amplifiers. The operating characteristics of several different types of gallium-arsenide reactive diodes are discussed and mention is made of simple design formulas which may be used to tentatively evaluate the performance to be expected from such diodes. Noise figure measurements are included in a résumé covering some of the experimental results that have been obtained using gallium-arsenide point-contact diodes as variable reactance elements in microwave parametric amplifiers.

Characterization of Microwave Variable Capacitance Diodes—Sverre T. Eng (p. 11)

This paper will describe the electrical characterization of microwave variable capacitance diodes. The importance of some of the diode parameters is discussed from the application point of view, and suitable measurement techniques for these parameters are described, together with actual measurement data on some diodes.

First, a general four-terminal transformation method is used, and some approximations lead to a fairly easy and accurate method of studying device characteristics. A resonant-cavity method is also considered, and it is explained under what condition it leads to a very simple test of the diode Q.

Finally, a method is presented which is based upon modifications of the Weissfloch canonical network. These simplifications can be used to get an easy interpretation of the junction impedance or the diode Q.

A Study of the Optimum Design of Wide-Band Parametric Amplifiers and Up-Converters—George L. Matthaei (p. 23)

Single-diode parametric amplifiers or up-converters using multiple-resonator filters as coupling networks can be made to have considerably larger bandwidths than corresponding amplifiers having single-resonator coupling circuits. Data are presented from which the coupling-filter bandwidths required for given coupling network complexity, diode parameters, and required gain can be determined for both parametric amplifiers and up-converters, In the cases of nondegenerate parametric amplifiers and up-converters, the fact that the diode must be brought to resonance at more than one frequency has an added limiting effect on bandwidth. Some trial amplifier designs are shown, and important considerations in the synthesis of the coupling filters are noted. It is seen that for the case of upper-sideband up-converters, if a filter having n resonators is used in both the input and upper-sideband circuits, then the over-all response can be made to correspond to that of a filter with $2n$ resonators. The gain characteristics of the trial amplifier designs as determined with a digital computer are included. Computed responses ranging in bandwidth from 9 to 27 per cent are obtained for multi-resonator designs having $C_1/C_0=0.25$.

A Low-Noise X-Band Parametric Amplifier Using a Silicon Mesa Diode—R. D. Weglein and F. Keywell (p. 39)

This paper summarizes a cooperative effort to develop silicon mesa variable-capacitance diodes and to evaluate their potential for achieving low-noise amplification in the high microwave frequency range. Cutoff frequencies of about 70 kMc at zero-bias voltage (corresponding to 140 kMc at maximum reverse bias voltage) with a total permissible voltage swing in excess of 5 volts have been obtained.

A versatile degenerate X-band parametric amplifier was developed which, when used in conjunction with these silicon mesa diodes, achieved a radiometer noise temperature of 130°K at 8.5 kMc with a 50-Mc bandwidth at 17-db gain. The measured performance of the diode (figure of merit) is compared with the first-order theory in an operating radar system. The over-all performance of the amplifier improved the observed system sensitivity by 6 db.

Design and Operation of Four-Frequency Parametric Up-Converters—J. A. Luksch, E. W. Matthews, and G. A. VerWys (p. 44)

A theoretical analysis of a four-frequency parametric-diode up-converter is presented, retaining both sum and difference frequencies generated by mixing of pump and signal. Upper and lower sideband up-converters are compared, and it is shown that the gain limitations of the former can be overcome by combination with the latter, without appreciable loss of stability. Three different parametric amplifier configurations utilizing this four-frequency mode of operation have been designed, fabricated, and tested. These designs utilize sum-frequency up-conversion from 400 to 9400 Mc, and have exhibited noise figures below 1.5 db, gain in excess of 12 db, and bandwidths greater than 8 Mc.

Ferrites with Planar Anisotropy at Microwave Frequencies—Isidore Bady (p. 52)

Materials with an easy plane of magnetization (planar anisotropy) have recently been discovered. The large anisotropy field that tends to keep the magnetization in the easy plane reduces the field required to cause ferromagnetic resonance, which makes the material promising for microwave applications. Equations are derived for the susceptibility, taking into account losses and a finite medium. Propagation in a longitudinal and transverse static field is considered. The location of a slab in a rectangular waveguide for minimum loss in the forward direction, and the use of the material as a phase shifter, are discussed. Experimental microwave data on some materials are given, and also data on an isolator and phase shifter incorporating these materials.

W de-Band Resonance Isolator—W. W. Anderson and M. E. Hines (p. 63)

A parallel-plate transmission line loaded with capacitors or high dielectric constant material along a narrow strip has a circularly polarized RF magnetic field everywhere external to the loading over a very broad band. The magnetic resonance line of a narrow line-width ferrite was inhomogeneously broadened by a very inhomogeneous magnetic field to provide resonance absorption over a wide frequency range. A prototype structure has given better than 15 db per inch attenuation in the reverse direction over a bandwidth from 1.5 kMc to 6.0 kMc. The forward loss caused by the ferrite is about 0.2 db to 0.4 db over this range of operation.

Maser Operation at Signal Frequencies Higher than Pump Frequency—Frank R. Arams (p. 68)

Methods using harmonic spin coupling for operating solid-state masers with signal frequencies higher than the pump frequency are discussed. Expressions for the population inversion ratios are presented, and the maximum signal-to-pump-frequency ratios are calculated.

Experimental data is presented on a ruby maser which is operated using the symmetrical method. Amplification was obtained at signal frequencies from 10,320 to 10,740 Mc, using pump frequencies ranging from 9580 to 9670 Mc.

An experiment in which maser operation is obtained simultaneously at two frequencies is described.

A Solid-State Microwave Source from Reactance Diode Harmonic Generators—T. M. Hyltin and K. L. Kotzebue (p. 73)

The generation of harmonics with reasonable efficiencies has been made possible by the application of high Q nonlinear reactance diodes. An approximate solution for the conversion loss of harmonic generators utilizing these devices has been obtained and design curves relating conversion loss with harmonic number, diode Q, and voltage-capacitance coefficient are presented. Harmonic generators have been operated with silicon and gallium-arsenide mesa diodes in the UHF region and conversion losses approaching the theoretical value have been obtained. Three harmonic generation stages in miniature modular packages were cascaded to obtained 7-mw output at S band. These stages were driven by a transistorized crystal-controlled oscillator and power amplifier which supplied 200 mw at 140 Mc from 1.3-w dc.

Propagation of Waves in a Plasma in a Magnetic Field—William P. Allis (p. 79)

The propagation of electromagnetic waves in a plasma in a magnetic field as given by the

Appleton-Hartree theory is discussed in terms of the wave normal surfaces instead of the more conventional propagation vector plots, and the "ordinary" and "extraordinary" waves are defined in terms of their polarizations instead of using a continuity argument. This gives a different picture of "a wave" which has some advantages. In particular, "whistlers" become obvious, as are regions of high reflection and high absorption.

The Appleton-Hartree theory is then extended to include the effect of electron temperature, and this results in a third wave whose velocity is of the order of electron thermal motions.

Magnetoplasma Effects in Solids—Benjamin Lax (p. 83)

Plasmas in solids show a more complex behavior than in gases since they reflect the symmetry properties of crystals. Since the carrier concentration has a wide range in semiconductors and metals, the plasma phenomena can be studied from microwaves to the ultraviolet. The effect of magnetic fields on the electromagnetic properties of plasmas has been experimentally investigated at microwave and infrared frequencies and has been utilized to measure dielectric constant and band structure of such solids in the limit of low magnetic fields. The magneto-plasma exhibits effects analogous to the galvanomagnetic phenomena. However, near resonance in the classical limit, they show up as depolarizing effects in semiconductors and also give rise to a new type of cyclotron resonance under anomalous skin conditions in metals.

Coherent Excitation of Plasma Oscillations in Solids—David Pines (p. 89)

Considerations are put forth concerning the feasibility of observing the coherent excitation of plasma oscillations in a two-component plasma of electrons and holes in semiconductors or semimetals. By coherent excitation is meant the onset of a high-frequency ("two-stream") instability arising from an appreciable drift of electrons vs holes under the action of an applied electric field. Conditions favorable to coherent excitation include a sizeable difference in electron and hole masses, and long relaxation times for both kinds of particles. The extent to which such conditions are present in InSb is discussed.

Pulsed Millimeter-Wave Generation Using Ferrites—B. J. Elliott, T. Schaug-Pettersen and H. J. Shaw (p. 92)

A method is described for generating pulsed RF energy in the millimeter-wave spectrum. Low-loss garnets are used in the uniform precessional mode to store energy at *S* band and radiate at a higher frequency, which is controlled by the total magnetic field. Details are given of a *K*-band generator which operates at frequencies up to 32 kMc.

Correspondence (p. 95)

PGMTT News, 1960 National Microwave Symposium (p. 103)

Contributors (p. 105)

Space Electronics and Telemetry

Vol. SET-7, No. 1, March, 1961

Chairman's Message—R. V. Werner (p. 1)

Frontispiece (p. 2)

On Coded Phase-Coherent Communications—A. J. Viterbi (p. 3)

The merits of phase-coherent communications are widely recognized for both discrete and continuous modulation systems. The relative performances of phase-coherent and noncoherent transmission of binary data in the presence of additive white Gaussian noise have been analyzed and compared. This paper considers the result of encoding independent equiprobable binary words or sequences of independent binary digits into sets of binary code words. These are transmitted over a channel perturbed by additive white Gaussian noise and detected by correlating them with their stored or locally generated replicas at the receiver.

The word error probabilities and bit error probabilities for low cross-correlation codes are determined as a function of the ratio

$$\frac{\text{(received signal energy)/bit}}{\text{(noise power)/(unit bandwidth)}}$$

The received information rate and the potential channel capacity are also computed. It is shown that in the limit as the code word length and the bandwidth approach infinity, the received information rate approaches the channel capacity for only one value of the above ratio.

Data Transmission for the NRL Space Surveillance System—M. G. Kaufman and F. X. Downey (p. 15)

A data-transmission system has been developed which links four distant receiving sites of the U. S. Navy Space Surveillance system to a data-reduction center located at Dahlgren, Va. The receiving sites form a fence located on a great circle route across the southern United States from Georgia to California. Each receiver site is coupled to the data center by a commercial, voice-quality, duplex (two-way) telephone line. Standard FM telemetry techniques are used to transmit eight channels of analog data on each telephone line. These data are transm tted on eight discrete frequency-modulated carriers in a frequency band from 270 cps through 2455 cps. In addition to these FM data carriers, unmodulated tones are used for monitoring, compensation, and command functions.

The data from each receiver site are permanently stored on paper recordings at the data-reduction center, so that this information can be assimilated at one location on a real-time basis. These data are used to compute the orbital parameters of satellites detected by the Space Surveillance system.

The data-transmission system has been in operation for a year on a 24-hour basis with negligible down time. Off-line calibration techniques have been employed so that errors introduced into the data by the transmission system can be held to 2 per cent without interfering with the detection capabilities of the surveillance system. Tests indicate that the number of channels can be increased from 8 to 24 per telephone line by the use of crystal-controlled oscillators and crystal filters.

The Role of the Jet Propulsion Laboratory in Project Echo—W. K. Victor and R. Stevens (p. 20)

The Jet Propulsion Laboratory of the California Institute of Technology was responsible for the West Coast Transmit/Receiver Station which was used in Project Echo for communicating with the East Coast Station of the Bell Telephone Laboratories utilizing the radio reflection properties of the satellite balloon. This report describes the work performed by JPL and presents some of the results of the Echo experiment.

Contributors (p. 29)

Ultrasonics Engineering

Vol. UE-8, No. 1, March, 1961

Application of Ultrasonic Light Modulation to Signal Recording, Display, Analysis and Communication—A. H. Rosenthal (p. 1)

Ultrasonic Doppler for Distance Measurement—Melvin Wachspress (p. 6)

High-efficiency transducers are utilized in a Janus Doppler system for distance measurement. Equations for the Janus system are developed for sound waves in a moving medium. A lightweight sonic Doppler system is described which is comparable in size and weight to a modern auto radio. An experimental Janus system is discussed. Spectrum photographs for various road surfaces are presented.

A High-Efficiency Transducer for Transmission to Air—Jack Kritz (p. 14)

The basic problems of transduction to low acoustic impedance media are considered. The advantages offered by flexure mode vibration of driving elements are shown. Selection of a driving element consisting of a free disk supported at the nodal circle is made. Fabrication in quartz and ceramic is described. Test results on transducers in the range 5 kc to 200 kc are given with transducer efficiencies of better than 90 per cent obtained. Transducer mounting in parabolic reflectors and the resultant beam patterns are shown.

Nondestructive Measurement of Tensile and Compressive Stresses—Rabah A. Shahbender (p. 19)

This paper discusses the experimental results of an investigation to determine the effects of stresses on the velocity of propagation of ultrasonic waves. The test data indicate measurable changes in the velocities of shear waves propagating transverse to the direction of applied stress.

On the basis of these data, a circularly polarized shear wave and a longitudinal wave are propagated transverse to the stress to determine the stress, the change in the polarization of the shear wave per unit distance being a measure of the stress.

A special combination of longitudinal and shear mode transducers is utilized to generate and detect the required wave.

The Ultrasonically-Coupled Oscillator—Yujiro Yamamoto (p. 22)

In this paper the theory of an oscillator consisting of an amplifier incorporating an ultrasonic feedback-path is presented. In the course of study, two methods of controlling the oscillation of the ultrasonically-coupled oscillator are investigated: 1) a highly selective circuit in the electronic amplifier controls the oscillation, and 2) the effective length of the feedback path is employed as a frequency-selective element of the oscillator.

Properties of Lamb Waves Relevant to the Ultrasonic Inspection of Thin Plates—T. N. Grigsby and E. J. Tajchman (p. 26)

An experiment is described, using $\frac{1}{32}$-inch thick steel plates, in which the insertion loss of artificial flaws was measured. The artificial flaws consisted of $\frac{1}{32}$-inch wide saw cuts of depths varying from 0.000 inch to 0.021 inch in increments of 0.003 inch. No simple relationship was found relating the insertion loss to the depth of the saw cuts, although the presence of a flaw was usually indicated. Measurements of the group velocity of Lamb waves are described and the values obtained are shown to agree with calculated values. The theory of Lamb waves is discussed, and the results of computation of the frequency equation for the symmetrical case for a ratio of longitudinal velocity to shear velocity of 1.8 are given in the form of curves of phase velocity vs frequency-thickness product. Curves of group velocity vs frequency-thickness product for this case are also given. A formula is derived which gives group velocity as a function of phase velocity and frequency-thickness product. A derivation of the frequency equation is given.

Contributors (p. 34)

Abstracts and References

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

UDC NUMBERS

Certain changes and extensions in UDC numbers, as published in PE Notes up to and including PE 666, will be introduced in this and subsequent issues. The main changes are:

Artificial satellites:	551.507.362.2	(PE 657)
Semiconductor devices:	621.382	(PE 657)
Velocity-control tubes, klystrons, etc.:	621.385.6	(PE 634)
Quality of received signal, propagation conditions, etc.:	621.391.8	(PE 651)
Color television:	621.397.132	(PE 650)

The "Extensions and Corrections to the UDC," Ser. 3, No. 6, August, 1959, contains details of PE Notes 598–658. This and other UDC publications, including individual PE Notes, are obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W. 1, England.

The Index to the Abstracts and References published in the PROC. IRE from February, 1959 through January, 1960 is published by the PROC. IRE, June, 1960, Part II. It is also published by *Electronic Technology* (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*) and included in the April, 1960 issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.232–14 **1055**

Bubble Transducer for Radiation High-Power Low-Frequency Sound in Water—C. C. Sims. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1305–1308; October, 1960.) "The bubble transducer consists of a conventional low-impedance electrodynamic driver with an air-filled rubber membrane, or 'bubble,' of adjustable volume over the diaphragm. The system at resonance produces large volume displacement of the water with small linear diaphragm displacement at frequencies below 200 c/s. For constant diaphragm volume velocity, the addition of the resonating bubble increases the power output by a factor of approximately 1000."

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears each June and December at the end of the Abstracts and References section.

534.283–8:538.6 **1056**

On the Theory of Ultrasonic Absorption of Metals in a Strong Magnetic Field: Part 1—E. A. Kaner. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 212–218; January, 1960.) The absorption coefficient is calculated considering closed electron orbits and an arbitrary law of electron dispersion.

534.614–8 **1057**

Determination of Ultrasonic Velocities by Measurement of Angles of Total Reflection—W. E. Mayer. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1213–1215; October, 1960.) The method is based on the total reflection of a sound beam from a liquid/solid boundary. The velocity of longitudinal and shear waves in the solid are determined by locating the angles of maximum reflection in the liquid.

534.75 **1058**

On the Pitch of Periodic Pulses—J. L. Flanagan and N. Guttman. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1308–1319; October, 1960). "Subjects adjusted the frequency of one periodic pulse train to match the pitch of another train fixed in frequency. Two modes of pitch perception are found. In the first mode, for pulse rates less than 100/sec, the pulse trains are ascribed a pitch equal to the number of pulses per second, regardless of the polarity pattern of the pulses. In the second mode, for fundamental frequencies in excess of 200 c/s, the sounds are assigned a pitch equal to the fundamental frequency."

534.75 **1059**

Pitch of Periodic Pulses without Fundamental Component—J. L. Flanagan and N. Guttman. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1319–1328; October, 1960.) See 1058 above. With filtered pulses a third mode of pitch perception is identified at fundamental frequencies above 1 kc.

534.75 **1060**

Monaural Temporal Masking Investigated by Binaural Interaction—N. Guttman, W. A. van Bergeijk, and E. E. David, Jr. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1329–1336; October, 1960.)

534.78:534.44 **1061**

A Versatile Method for Short-Term Spectrum Analysis in "Real-Time"—J. S. Gill. (*Nature*, vol. 189, pp. 117–119; January 14, 1961.) A method is outlined for the spectrum analysis of speech as it proceeds in time. Examples are given of continuous spectrograms obtained in an analysis covering a frequency band of 3.3 kc, with different analyzer bandwidths down to 50 cps.

534.9 **1062**

Binaural Summation of Loudness—G. S. Reynolds and S. S. Stevens. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1337–1344; October, 1960.)

681.84 **1063**

Recording and Reproducing Equalization for Coarse- and Fine-Groove Records—P. J. Guy. (*Electronic Engrg.*, vol. 33, pp. 9–15; January, 1961.) The definitions of the standard characteristics are explained and the effects of generator and load impedances are discussed.

ANTENNAS AND TRANSMISSION LINES

621.315.212:621.372.51 **1064**

A Convenient Transformer for Matching Coaxial Lines—B. Bramham. (*Electronic Engrg.*, vol. 33, pp. 42–44, January, 1961.) Two short sections of the lines to be joined are connected between the lines in such a way that the three impedance changes cancel at a particular frequency.

621.372.2:621.372.832.43 **1065**

Recent Advances in the Use of Coupled Transmission Lines as Directional Couplers—R. Koike. (*Proc. IEE*, vol. 108, pt. B, pp. 120–124; January, 1961.) New designs of slot-coupled coaxial lines in which the slot spacing can be varied have been developed as power monitors and variable attenuators at frequencies up to 7 Gc.

621.372.22.011.21 **1066**

Calculations on Lines with Continuously Variable Impedance—A. V. J. Martin and F. J. Young. (*J. Phys. Radium*, vol. 19, suppl. to no. 7, *Phys. Appl.*, pp. 65A–70A; July, 1958.) Two methods are proposed for calculating the impedance of a tapered transmission line: a) by solution of differential equations, giving the impedance as a function of line length for a fixed frequency; b) by successive approximations, considering the line as a series of short sections, each having a constant impedance.

621.372.8.049 1067
Waveguide Techniques—O. Henke and G. Stricker. (*Frequenz*, vol. 14, pp. 94–104; March, 1960.) Various techniques of manufacturing waveguides are discussed in relation to size and accuracy requirements as dictated by the type of application and operating conditions.

621.372.825 1068
Characteristics of Ridge Waveguides—F. Young and J. Hohmann. (*Appl. Sci. Res.*, vol. B8, no. 4, pp. 321–336; 1960.) General expressions for the parameters of single- and double-ridged waveguides are derived. Detailed curves of their values are given for waveguides having width/height ratios of 1:1 and 2:1.

621.372.832.8 1069
The Microwave Circulator—E. Pivit and W. Stösser. (*Frequenz*, vol. 14, pp. 77–84; March, 1960.) The operating principles of the various types of circulator are summarized, and the theory of the phase-shift type is considered in detail in order to evaluate the tolerances of the constituent elements. Design data and characteristics of phase-shift circulators are given for various frequency ranges.

621.396.674–428 1070
Spiral Antennas—W. L. Curtis. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 298–306; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1512; August, 1960.)

621.396.674–428 1071
The Archimedean Two-Wire Spiral Antenna—J. A. Kaiser. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 312–323; May, 1960.) Abstract, Proc. IRE, vol. 48, p. 1512; August, 1960.)

621.396.674.3 1072
Transient Phenomena associated with Sommerfeld's Horizontal Dipole Problem—H. J. Frankena. (*Appl. Sci. Res.*, vol. B8, no. 4, pp. 357–368; 1960.) A theoretical treatment gives expressions for the traveling wave from a horizontal dipole having a moment varying arbitrarily with time and situated above a dielectric interface.

621.396.674.3 1073
Radiation of Pulses Generated by a Vertical Electric Dipole above a Plane, Nonconducting Earth—A. T. de Hoop and H. J. Frankena. (*Appl. Sci. Res.*, vol. B8, no. 4, pp. 369–377; 1960.) An expression is obtained for the EM field in air. The reflected-wave portion is expressed as an integral which can be evaluated numerically.

621.396.677:621.397.61 1074
The Use of a High-Gain Television Transmitting Aerial in a Populous Area—G. D. Monteath, G. H. Millard, and D. J. Whythe. (*Proc. IEE*, vol. 108, pt. B, pp. 65–74; January, 1961.) A review is given of the problems encountered in providing a satisfactory service using an eight-tier antenna in an urban area, with particular reference to the Crystal Palace station.

621.396.677.3 1075
Determination of the Gain and Directional Characteristics of a Yagi Aerial for Decimetre Waves—D. Stahl. (*Rundfunktech. Mitt.*, vol. 4, pp. 85–87; April, 1960.) The optimum arrangement of transmitting and receiving antennas for the measurement of gain and the determination of polar diagrams is shown and the results of measurements on four Yagi antennas for use at 569 Mc are given.

621.396.677.3 1076
A New Mathematical Approach for Linear Array Analysis—D. K. Cheng and M. T. Ma. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 255–259; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1511; August, 1960.)

621.396.677.3 1077
Log Periodic Dipole Arrays—D. E. Isbell. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 260–267; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1511; August, 1960.)

621.396.677.3:523.164 1078
The Crossed-Grating Interferometer: a New High-Resolution Radio Telescope—W. N. Christiansen, N. R. Labrum, K. R. McAlister, and D. S. Mathewson. (*Proc. IEE*,, vol. 108, pt. B, pp. 48–58; January, 1961.) A description is given of a radio telescope, operating at 21-cm λ, with two ground arrays each consisting of 32 paraboloid antennas spaced uniformly along a 1200-foot baseline. Details are given of its operation in measuring the brightness distribution of the solar disk.

621.396.677.3:621.396.677.71 1079
Mutual Coupling Effects in Large Antenna Array: Part 1—Slot Arrays—S. Edelberg and A. A. Oliner (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 286 297; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1512; August, 1960.)

621.396.677.3:621.396.965 1080
Mutual-Impedance Effects in Large Beam Scanning Arrays—P. S. Carter, Jr. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 276–285; May, 1960. Abstract, Proc. IRE, vol. 48, pp. 1511–1512; August, 1960.)

621.396.677.4 1081
Beam Pointing Errors of Long Line Sources—M. Leichter. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 268–275; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1511; August, 1960.)

621.396.677.75 1082
Coupled Leaky Waveguides: Part 1—Two Parallel Slits in a Plane—S. Nishida. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 323–330; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1512; August, 1960.)

621.396.677.83 1083
Experimental Study of a Diffraction Reflector—J. H. Provencher. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 331–336; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1512; August, 1960.)

621.396.677.833:523.164 1084
Malvern 45-ft Radio Telescope—J. S. Hey and V. A. Hughes. (*J. IEE*, vol. 7, pp. 7–11; January, 1961.) The precision of construction of the telescope for cm-λ observations is discussed and radio-astronomy applications are outlined.

621.396.677.833.2 1085
On the Beam Deviation Factor of a Parabolic Reflector—Y. T. Lo. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 347–349; May, 1960.)

AUTOMATIC COMPUTERS

681.142 1086
Computer-Generated Displays—R. T. Loewe, R. L. Sisson, and P. Horowitz. (Proc. IRE, vol. 49, pp. 185–195; January, 1961.) A survey of methods and techniques in the presentation of computer outputs is given. 52 references.

681.142:538.221:539.23 1087
Magnetic Film Memory Design—J. I. Raffel, T. S. Crowther, A. H. Anderson, and T. O. Herndon. (Proc. IRE, vol. 49, pp. 155–164; January, 1961.) A storage system for 32 ten-bit words has been operated successfully in the destructive mode with a minimum cycle time of 0.4 μsec.

681.142:621.3.087.4:621.395.625.3 1088
High-Density Digital Magnetic Recording Techniques—Hoagland and Bacon. (See 1316.)

681.142:621.318.4 1089
The Simulation of Neural Elements by Electrical Networks Based on Multi-aperture Magnetic Cores—A. E. Brain. (Proc. IRE, vol. 49, pp. 49–52; January, 1961.) The basic operations of multilevel storage, gating, controlling threshold level, and summation are common to many simulating networks and can be attained by use of multi-aperture magnetic cores.

681.142:621.318.57 1090
New Storage Systems based on the Magnetization of Three-Dimensional Elements—R. Vacca. (*Ricerca Sci.*, vol. 30, pp. 445–459; April, 1960.) Three systems are described: two are diode-controlled circuits for switching magnetic storage devices, and the third is Biax, a multi-aperture ferrite element using the principle of flux interference. See also 1959 IRE WESCON Convention Record, vol. 3, pt. 4, pp. 40–54 (C. L. Wanlass and S. D. Wanlass).

681.142:621.318.57:621.372.44 1091
Nonstationary Parametrons—Onose. (See 1098.)

681.142:621.373.029.6:621.372.44 1092
A Computer Subsystem using Kilomegacycle Subharmonic Oscillators—I. Abeyta, F. Borgini, and D. R. Crosby. (Proc. IRE, vol. 49, pp. 128–135; January, 1961.) A phase-locked oscillator is described which operates at 1.85 Gc and has a logical operation time of 11 nsec.

681.142:621.374.4:621.382.23 1093
Frequency Division by Carrier Storage—Ryan. (See 1107.)

681.142:621.374.5 1094
Computer Memories: a Survey of the State-of-the-Art—J. A. Rajchman. (Proc. IRE, vol. 49, pp. 104–127; January, 1961.)

681.142:621.382.23 1095
A Survey of Tunnel-Diode Digital Techniques—R. C. Sims, E. R. Beck, Jr., and V. C. Kamm. (Proc. IRE, vol. 49, pp. 136–146; January, 1961.)

681.142:621.382.23 1096
Calculated Waveforms for Tunnel-Diode Locked Pair—H. R. Kaupp and D. R. Crosby. (Proc. IRE, vol. 49, pp. 146–154; January, 1961.) The phase-locked pair functions, as a computer element, in a similar manner to that of a phase-locked harmonic oscillator, and also requires a three-phase power supply. Its main advantages are high gain and high transistion speed.

CIRCUITS AND CIRCUIT ELEMENTS

621.3.011.22:621.38 1097
The Negative-Resistance Concept in Modern Electronics—G. D. Sims and I. M. Stephenson. (*J. Electronics and Control*, vol. 9, pp. 349–383; November, 1960.) A survey with particular reference to the operation of semiconductor diodes and microwave tubes is given.

621.318.57:621.372.44:681.142 1098

Nonstationary Parametrons—K. Onose. (*Rev. Elec. Commun. Lab., Japan*, vol. 8, pp. 204–210; May/June, 1960.) To eliminate the necessity for the critical adjustment of amplitude and phase of the excitation current, dc-controlled and diode-controlled parametrons were investigated. A fast decay time was achieved with diode control, resulting in a high speed of operation in logic circuits.

621.318.57:621.382.333 1099

The Cryosistor—a New Low-Temperature Three-Terminal Switch—I. Melngailis. (PROC. IRE, vol. 49, pp. 352–354; January, 1961.) A reverse-biased *p-n* junction is used to control the ionization between two ohmic contacts. Its uses might be a) as a fast monostable or bistable switch, and b) as a pulse amplifier.

621.372.41 1100

Observations on some Theorems for Linear Passive Networks—F. Gasparini. (*Alta Frequenza*, vol. 29, pp. 90–95; February, 1960.) An extension of Cohn's theorem for linear passive two-poles [see *e.g.*, 3056 of 1957 (Vratsanos)] is proposed, facilitating the determination of energy relations in such networks.

621.372.413 1101

Theory of Forced Oscillations in Electromagnetic Cavities, and the Equivalent Circuit of a Cavity 2-n-Pole—H. J. Butterweck. (*Arch. elekt. Übertragung*, vol. 14, pp. 101–114; March, 1960.) The EM field inside an excited cavity coupled to a two-way transmission line is calculated on the basis of a series expansion of arbitrary vector fields in cavity resonators in terms of orthogonal eigenfunctions. The equivalent circuit derived consists of an infinite number of lumped elements; approximation formulas for practical application are given.

621.372.413:537.311.33 1102

The Cylindrical TE_{11}-Mode Cavity including a Semiconductor Sample with Tensorial Electrical Conductivity—N. Watanabe. (*Rev. Elec. Commun. Lab., Japan*, vol. 8, pp. 256–265; May/June, 1960.) The impedance matrix and characteristic *S*-matrix are derived for two configurations of the input guides. One case is suitable for the measurement of Hall mobility, and for the other the proper conditions for use as an isolator are investigated.

621.372.512 1103

Iterative Networks as Coupling Networks in Oscillator Circuits—R. Paul. (*Frequenz*, vol. 14, pp. 84–94; March, 1960.) The conditions for self-oscillation are established for tube and transistor oscillators. The characteristics of iterative networks consisting of purely resistive and reactive elements are summarized, and their application in tube and transistor oscillators is investigated. Formulas are given for operating parameters and are verified experimentally.

621.372.543.2 1104

Explicit Formulae for the Calculation of the Characteristic Function of Filters with Tchebycheff Pass-Band Behaviour—A Fettweis. (*Rev. HF, Brussels*, vol. 4, no. 12, pp. 263–271; 1960. In English.)

621.373.421.11 1105

A Wide-Band Voltage-Controlled Oscillator—M. A. Weston. (*Electronic Engrg.*, vol. 33, pp. 2–5; January, 1961.) Continuously variable frequency control of an LC oscillator over the range 1.5–3.0 Mc is obtained by electronic switching of a number of fixed capacitors across the inductance, the number depending on a dc control signal voltage.

621.373.431.2:621.382.3 1106

A Transistor Relaxation Oscillator and its Applications—E. Bächle. (*Elektronik*, vol. 9, pp. 68–72; March, 1960.) The modifications of the Schmitt trigger circuit necessitated by the use of transistors are investigated; practical circuits and operating characteristics are given.

621.374.4:621.382.23:681.142 1107

Frequency Division by Carrier Storage—W. D. Ryan. (*Electronic Engrg.*, vol. 33, pp. 40–41; January, 1961.) A circuit using a junction diode with forward bias that produces output pulses at half the frequency of a 50-kc input sinusoidal voltage is described. Possible computer applications are noted.

621.374.4+621.376.233].029.65 1108

A Harmonic Generator and Detector for the Short-Millimetre-Wave Region—H. W. de Wijn. (*Appl. Sci. Res.*, vol. B8, no. 4, pp. 261–264; 1960.) The construction of an instrument of the open-crystal type is described.

621.375.9:621.372.44 1109

Amplification by means of Nonlinear Reactances—G. B. Stracca and A. Butti. (*Ricerca Sci.*, vol. 30, pp. 313–343; March, 1960.) The theory of parametric amplification is considered on the basis of the Manley-Rowe power relations. The properties of varactors and their use in parametric amplifiers are discussed, and some experimental results are given.

621.375.9:621.372.44 1110

Experimental Demonstration of Parametric Amplification—E. Meyer and W. Eisenmenger. (*Z. Phys.*, vol. 158, pp. 379–385; April 4, 1960.) Simple experiments are described to demonstrate the basic principles of parametric amplification, a) using a variable-capacitance circuit, and b) by means of a mechanical system for setting up waves on a water surface.

621.375.9:621.372.44:621.372.632 1111

Noise Figure and Gain of Parametric Converters—H. Heffner and G. Wade. (*J. Appl. Phys.*, vol. 31, p. 2316; December, 1960.) A correction is given for a term in a noise-figure equation for a parametric frequency converter appearing in an earlier communication (77 of 1959).

621.375.9:621.372.44:621.385.6 1112

An Electron-Beam Parametric Amplifier for the 200-Mc/s Region—Chalk. (See 1370.)

621.375.9:621.383:535.376 1113

Power Amplifiers based on Electro-optical Effects; a Survey—G. Diemer and J. G. van Santen. (*Philips Res. Repts.*, vol. 15, pp. 368–389; August, 1960.) The gain, bandwidth, and noise factor of electroluminescent-photoconductive transducers are considered, taking into account the effects of feedback and matching. Suggested practical applications of single and cascade devices are discussed. See also 442 of 1960 (Lochinger and Strutt).

621.375.9.029.64:537.311.33 1114

Instability Effects in a Semiconductor Amplifier with Negative Effective Carrier Mass—A. L. Zakarov. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 665–667; February, 1960.) A critical comment on the proposal of Krömer (1824 of 1959) is given.

621.376.22:621.374.43 1115

Simple Periodic Behaviour of the Ring Modulator—V. Belevitch, J. Meinquet, and P. van Bastelaer. (*Rev. HF, Brussels*, vol. 4, no. 12, pp. 272–277; 1960.) An analysis is made of ring-modulator theory where the signal and carrier frequencies are equal, or one is a multiple of the other. Curves are given for determining the performance of frequency dividers of the Miller type.

GENERAL PHYSICS

535.325:535.14 1116

Theory of the Complex Refractive Index—C. A. Mead. (*Phys. Rev.*, vol. 120, pp. 854–859; November 1, 1960.) A quantum-mechanical formalism is developed for representing the complex refractive index of a gas. See also 2690 of 1958.

537.311.1 1117

Theory of Impurity Resistance in Metals—J. S. Langer. (*Phys. Rev.*, vol. 120, pp. 714–725; November 1, 1960.) A many-body technique is developed for the calculation of the dc resistivity of a Fermi fluid in the presence of a few randomly-distributed fixed impurities.

537.311.62:538.569.4 1118

Electron Relaxation Time in a High-Frequency Electromagnetic Field and the Surface Impedance of a Metal—R. N. Gurzhi and M. Ya. Azbel'. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 524–528; February, 1960.) An analysis of effects due to quantization of the EM field and the electron orbits in a constant magnetic field is given.

537.533 1119

The Passage of Electrons through a Rectangular Potential Barrier with a Small Cylindrical Inhomogeneity—N. B. Aĭzenberg. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1178–1185; June, 1960.) A treatment is made of the tunnel effect in electron emission by first-order perturbation theory.

537.56 1120

On Certain Special Features of Ohmia Heating of an Electron Gas in a Plasma—A. V. Gurevich. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 116–121; January, 1960.) An investigation of ohmic heating in a constant field, taking account of inelastic collisions, shows that the electron temperature may be in a steady state only for low values of electric field intensity. See also 2532 of 1959.

537.56 1121

The Origin of a Backward Wave in a Nonmagnetized Plasma Cylinder Bounded by Air—W. O. Schumann. (*Z. angew. Phys.*, vol. 12, pp. 145–148; April, 1960.) The conditions for the existence of such a wave are derived.

537.56:538.56 1122

Radiation of Plasma Oscillation—S. Kojima and S. Hagiwara. (*J. Phys. Soc. Japan*, vol. 15, p. 1904; October, 1960.) Radiation of EM waves from a mercury discharge tube was picked up by a horn feeding a receiver tuned to 9600 Mc. A peak value of received signal was obtained for a discharge current corresponding to a plasma resonant at 9600 Mc. Reflections from the plasma were also observed; they reached a peak value at the same discharge current, but did not fall with higher currents. See also 1182 of 1960 (Kojima *et al.*).

537.56:538.56 1123

A Note on 'Effects of Relatively Strong Fields on the Propagation of E.M. Waves through a Hypersonically Produced Plasma'—J. M. Fiskin and W. B. Sisco. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 338–339; May, 1960.) See also 454 of 1960 (Sisco and Fiskin).

537.56:538.56 1124

A Note on the Relation between the 'Exact' and 'Simplified' Theories for E.M. Wave Propagation in Ionized Gases—J. M. Anderson. (IRE TRANS. ON ANTENNAS AND PROPAGATION, vol. AP-8, pp. 337–338; May, 1960.) Comment

on, and correction to, the paper by Sisco and Fiskin (454 of 1960) is given.

537.56:538.569.4 1125
Cyclotron Absorption of Electromagnetic Waves in a Plasma—K. N. Stepanov. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 265–267; January, 1960.) A brief analysis of the absorption due to the thermal motion of electrons and ions is made.

537.56:538.63 1126
Plasma Oscillations in a Magnetic Field—I. Kaji and M. Kito. (*J. Phys. Soc. Japan*, vol. 15, pp. 1851–1861; October, 1960.) A theoretical treatment is made of longitudinal electron oscillations.

537.56:538.63 1127
Stability of a Plasma Sheet in Time-Periodic Magnetic Fields—L. Solymar. (*J. Electronics and Control*, vol. 9, pp. 391–396; November, 1960.) The behavior of a thin sheet of electrons in a time-varying square-wave magnetic field is investigated and it is shown that the growth of an initial disturbance in the first half-period is entirely annuled in the second, both the displacements and velocities returning to their initial values.

538.311:530.12 1128
The Field of an Electric Current in General Relativity—W. B. Bonnor. [*Proc. Phys. Soc.* (*London*), vol. 76, pp. 891–899; December, 1960.] A solution is given for a wire loop, for which it is necessary to endow the wire with a gravitational mass which corresponds to the gravitational energy of the current of the magnetic field created. The effect is not far below the present limit of experimental detection.

538.566:535.42 1129
Apparatus for Investigating Polarization in the Diffraction of Centimetric Electromagnetic Waves—A Mével and J. Mével. (*J. Phys. Radium*, vol. 19, suppl. to no. 12, *Phys. Appl.*, pp. 133A–139A; December, 1958.) A semi-automatic rotating analyzer for measurement of the polarization of scattered waves in free space is described. See also 1393 of 1958 (J. Mével).

538.566:535.42 1130
Solution of a Diffractive Problem—H. M. Nussenzveig. (*Phil. Trans. Roy. Soc. London. Ser. A*, vol. 252, pp. 1–51; October 15, 1959.) The problem of radiation from a semi-infinite parallel-plate waveguide, terminated by an infinite plane flange (double wedge) is investigated in the k-representation when the width of the double wedge is a) $\gg\lambda$, b) $\ll\lambda$.

538.566:535.43 1131
Theoretical and Experimental Study of Back-Scattering Cross-Section of an Infinite Ribbon—M. S. Macrakis. (*J. Appl. Phys.*, vol. 31, pp. 2261–2266; December, 1960.) "The geometrical-optics approximation is derived for the back-scattering cross-section per unit length of an infinite ribbon. A comparison is made with the exact theory, the approximate theory of Sommerfeld, the variation method, and with experimental results obtained through the space-separation method for the measurement of back-scattering cross-sections in a parallel-plate region."

538.566:537.56:621.375.9 1132
Reflection of Electromagnetic Waves from a Plasma Moving in Slow-Wave Waveguides—O. G. Zagoradnov, Ya. B. Faĭnberg, and A. M. Egorov. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 7–9; January, 1960.) Report of an experimental investigation carried out at 24.75 Mc, using a helix slow-wave structure to reduce the phase velocity of the waves to 1/200–1/375 c is given. The double Doppler effect observed in reflection from a plasma piston increased the frequency by 11–20 per cent. Application of the effect for microwave amplification and frequency multiplication is indicated.

538.569.2 1133
Possible Transmission of Electromagnetic Waves through a Metal in a Strong Magnetic Field—O. V. Konstantinov and V. I. Perel'. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 161–164; January, 1960.) Analysis shows that an EM wave propagating along a magnetic field can penetrate a metal plate perpendicular to the field if the Larmor frequency is higher than the frequency of the wave and much higher than the collision frequency, and if the Larmor electron radius is smaller than the wavelength in the metal.

538.569.4:535.33:621.375.9 1134
Microwave Spectroscopy—D. J. Millen. (*Proc. IEE*, vol. 108, pt. B, pp. 111–119; January, 1961.) The basis of molecular modulation is discussed with special reference to the Stark effect and maser amplification. The various types of spectrometer and their applications are discussed.

1135
Crossover Transistions—M. W. P. Strandberg. (*J. Phys. Chem. Solids*, vol. 16, pp. 39–43; November, 1960.) "Finite-rotation operators are used to simplify the calculation of second-order, crossover transitions for a spin 3/2 system with trigonal symmetry. The resonant frequency and transition probabilities for the $+3/2$, $-1/2$ crossover transition with negative D (or $+1/2$, $-3/2$ with positive D) are calculated."

538.569.4:538.222 1136
The Second Moment of a Paramagnetic Absorption Line taking account of the Effect of Fine and Hyperfine Structures—U. Kh. Kopvillem. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 151–156; January, 1960.) Formulas are derived for calculating the reduced second moment of a resonance line and the fine-structure constants of nuclear and electron paramagnetic absorption.

538.633 1137
Corbino Disk—D. A. Kleinman and A. L. Schawlow. (*J. Appl. Phys.*, vol. 31, pp. 2176–2187; December, 1960.) A comprehensive treatment is made of the behavior of the electric current and magnetic-field distribution patterns in a disk with concentric inner and outer contacts and a magnetic field applied along its axis. The effects discussed are large in materials with a high mobility, such as Bi, and the Corbino disk appears to be useful as a rectifier and voltage regulator.

539.2:537.122 1138
Spin Susceptibility of an Electron Gas—P. A. Wolff. (*Phys. Rev.*, vol. 120, pp. 814–819; November 1, 1960.) The generalized random-phase approximation is used to investigate the effects of electron-electron interaction on the spin susceptibility of all electron gas.

539.2:548.0 1139
Electron Levels in a One-Dimensional Random Lattice—H. L. Frisch and S. P. Lloyd. (*Phys. Rev.*, vol. 120, pp. 1175–1189; November 15, 1960.)

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.152.3:538.12 1140
Apparent Steady Component of the Interplanetary Magnetic Field—V. A. Bailey. (*Nature*, vol. 189, pp. 44–45; January 7, 1961.) Arguments based on the results of Coleman, *et al.* (see 3832 of 1960) and on the direction and sense of terrestrial storm fields show that the apparent steady component of the interplanetary field (also known as quiet-time field) is probably directed from the north to the south of the ecliptic.

523.16 1141
Existence of Net Electric Charges on Stars—L. Oster and K. W. Philip: V. A. Bailey. (*Nature*, vol. 189, pp. 43–44; January 7, 1961.) Comment is made on 2324 of 1960 and the author's reply is given.

523.164:551.510.535 1142
Cosmic Noise Absorption Measurements at Stanford, California, and Pullman, Washington—B. Lusignan. (*J. Geophys. Res.*, vol. 65, pp. 3895–3902; December, 1960.) The absorption was measured at 27.5 Mc using a riometer. After making allowances for the diurnal variations due to absorption of the D and F layers, a residual diurnal variation remains which has not yet been explained.

523.164:621.396.677.3 1143
The Crossed-Grating Interferometer: a new High-Resolution Radio Telescope—Christiansen, Labrum, McAlister, and Mathewson. (See 1078.)

523.164:621.396.677.833 1144
Malvern 45-ft Radio Telescope—Hey and Hughes. (See 1084.)

523.164.3 1145
A Survey of Radio Stars at a Frequency of 38 Mc/s—G. R. Whitfield. (*Mon. Not. R. Astr. Soc.*, vol. 120, no. 6, pp. 581–588; 1960.) "This paper describes a survey of small diameter radio stars made with an interferometer at 38 Mc/s. The positions and flux densities of 59 radio stars, all of which are confirmed by other surveys, are given in the catalogue."

523.164.3:523.75 1146
Relation of Jupiter's Radio Emission at Long Wavelengths to Solar Activity—J. W. Warwick. (*Science*, vol. 132, pp. 1250–1252; October 28, 1960.) Observations at Boulder during the period January–June, 1960, show a strong positive correlation between Jupiter's emission in the range 15–34 Mc and solar continuum emission. The time delay of one to two days indicates that fast solar particles at velocities of about 0.1 c may be involved in the planet's atmosphere or magnetic field. See also 523 of February (Carr, *et al.*).

523.164.32:551.510.535 1147
Solar Radio Emission on Centimetre Waves and Ionization of the E Layer of the Ionosphere—M. R. Kundu. (*J. Geophys. Res.*, vol. 65, pp. 3903–3907; December, 1960.) The correlation coefficient relating solar RF noise flux on wavelengths between 3 and 70 cm to daily and monthly E-layer indexes described by Minnis and Bazzard (3281 of 1959 and 1580 of 1960) has been calculated. The coefficient lies in the range 0.7–0.8 for wavelengths between 3 and 30 cm; above 50 cm it falls to 0.2.

523.165:551.507.362 1148
Radiation Environment in Space—H. E. Newell and J. E. Naugle. (*Science*, vol. 132, pp. 1465–1472; November 18, 1960.) A summary is given of information on radiations in space obtained by means of satellites and space probes.

523.165:551.507.362 1149
Relation between Results of Measurements of Charged Particles by means of Traps on Soviet Cosmic Rockets and Measurements of the Magnetic Field on the American Satel-

lite Explorer VI and the Rocket Pioneer V—K. I. Gringauz and S. M. Rytov. (*Dokl. Ak. Nauk SSSR*, vol. 135, pp. 48–51; November 1, 1960.) Results of measurements carried out by different methods on the Soviet rockets indicate the existence of a geomagnetic belt of charged particles 60,000 km from the center of the earth.

523.3:621.396.96 1150
Moon Reflection Studies with Bistatic Radar at 3000 Mc/s—A. W. Straiton and C. W. Tolbert. (*Commun. and Electronics*, pp. 436–440; September, 1960. Discussion.) The reception at Austin, Tex., of moon echoes of transmissions from Malvern, England, is reported. Discrete "spike" echoes persisted for about 0.25 second. Simultaneous observations at Malvern and Austin show no correlation between individual spikes. This is interpreted in terms of an undulating lunar surface.

523.53:621.396.96 1151
Radio Echo Measurements of the Orbits of Faint Sporadic Meteors—J. G. Davies and J. C. Hill (*Mon. Not. R. Astr. Soc.*, vol. 121, no. 5, pp. 437–462; 1960.) A survey has shown that many such meteors move in orbits of very short period. The more eccentric orbits are near the plane of the ecliptic but some are circular and have inclinations near 60°. The significance of the results is discussed.

523.53:621.396.96:535.42 1152
The Calculation of Meteor Velocities from Continuous-Wave Radio Diffraction Effects from Trails—J. S. Mainstone. (*Mon. Not. R. Astr. Soc.*, vol. 120, no. 6, pp. 517–529; 1960.) Meteor velocities are calculated from the time interval between cw echo maxima or minima. Neglect of the initial phase relation between echo and ground wave introduces large errors. An analytical method is described and preliminary results are given.

523.755 1153
The Structure and Magnetic Field of the Solar Corona—J. A. Högbom. (*Mon. Not. R. Astr. Soc.*, vol. 120, no. 6, pp. 530–539; 1960.) Radio observations in 1958 of the Crab nebula made at 38 Mc using four interferometers are described. Anisotropic spreading of the source occurred as it passed near the sun. Radial filamentary irregularities of ionization extending out to 30 solar radii are deduced.

550.385:539.16 1154
Possible Magnetic Effects with High-Altitude Explosions of Atomic Bombs—O. I. Leĭpunskiĭ. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 302–304; January, 1960.) The effects of a dense diamagnetic plasma expanding in the earth's magnetic field are discussed with reference to the periodic magnetic disturbances associated with the Argus experiment.

550.385.4:523.75 1155
An Approximate Method of Estimating the Size and Shape of the Stationary Hollow Carved Out in a Neutral Ionized Stream of Corpuscles Impinging on the Geomagnetic Field—V. C. A. Ferraro. (*J. Geophys. Res.*, vol. 65, pp. 3951–3953; December, 1960.) A two dimensional case is considered in which the permanent field is due to a current acting perpendicular to the stream. The breadth of the hollow at infinity is shown to be finite.

550.386 1156
The Latitudinal Distribution of Magnetic Activity in Canada—K. Whitham, E. I. Loomer, and E. R. Niblett. (*J. Geophys. Res.*, vol. 65, pp. 3961–3974; December, 1960.) Measurements of the hourly ranges in the principal horizontal field component at sixteen Canadian I.G.Y. stations are described. A narrow zone or area of enhanced magnetic activity exists at high latitudes with a maximum near Alert, $\phi=86°$. The seasonal and diurnal variations of magnetic activity at different latitudes are also considered.

551.507.362.2 1157
Perturbations of the Orbit of the Echo Balloon—I. J. Shapiro and H. M. Jones. (*Science*, vol. 132, pp. 1484–1486; November 18, 1960.) Predictions of the influence of solar radiation pressure are confirmed by the changes in orbit of satellite 1960ι during the first 12 days after its launching.

551.507.362.2 1158
Observed Solar Pressure Perturbations of Echo I—D. O. Muhleman, R. H. Hudson, D. B. Holdridge, R. L. Carpenter, and K. C. Oslund. (*Science*, vol. 132, p. 1487; November 18, 1960.) A decrease in perigee height of 3.0 km per day and an increase in the eccentricity of 0.00038 per day were observed during the first 10 days after launching of 1960ι. See also 1157 above.

551.507.362.2:621.317.361 1159
A Method for Interpreting the Doppler Curves of Artificial Satellites—G. Boudouris. (*J. Brit. IRE*, vol. 20, pp. 933–935; December, 1960.) A graphical-analytical method for determining the point of inflexion and the maximum slope is explained in detail. See also 1601 of 1960 (Boudouris, *et al.*).

551.507.362.2:621.391.812.3 1160
Irregular Fading of Satellite Transmissions—W. C. Bain. (*Nature*, vol. 189, p. 129; January 14, 1961.) The distribution of the mean scintillation index with magnetic inclination, derived from observations at Slough of 20-Mc transmissions from 1959ι, is discussed in relation to comments by Mawdsley (4215 of 1960) on a paper by Kent.

551.507.362.2:621.391.812.3 1161
Sudden Amplitude Variations of Sputnik III Signals—H. Whitney, H. Strick, J. Aarons, and J. Mott. (*J. Geophys. Res.*, vol. 65, pp. 4210–4212; December, 1960.) Two distinct causes of "drop-outs" in the signal received from 1958δ2 are identified: a) a change in transmitter characteristics, and b) a localized absorbing or scattering region in the ionosphere.

551.507.362.2:621.391.812.63 1162
Field-Aligned Ionospheric Irregularities and the Scintillation of Satellite Radio Transmissions—D. G. Singleton, G. J. E. Lynch, and J. A. Thomas. (*Nature*, vol. 189, pp. 30–31; January 7, 1961.) Observations of radio transmissions from 1959ι made at Brisbane during July–September, 1960 have been analyzed. Results indicate that the scintillation activity associated with propagation along the lines of the earth's magnetic field is more pronounced than that associated with other directions of propagation. The results are in agreement with those of Rush and Colin (PROC. IRE, vol. 46, pp. 356–357; January, 1958.).

551.507.362.2:621.396.96 1163
Radar Echoes obtained from Earth Satellites 1957 Alpha and 1957 Beta—W. E. Jaye, R. B. Dyce, and R. L. Leadabrand. (*Planet. Space Sci.*, vol. 5, pp. 50–58; January, 1961.) Observations at Stanford, Calif., using a 106-Mc radar with a 61-foot steerable paraboloid are described. Echoes from 1957α and β were seen on fifty-four occasions. The echoes usually lasted about 10 seconds as the satellite crossed the antenna beam. Cyclic fading at 0.1–0.9 cps occurred, usually with constant polarization, attributable to satellite tumbling. On one occasion 1957β passed within 200 km of an auroral disturbance detected on the radar; large amplitude fluctuations were seen and the echo lasted longer than usual.

551.510.535 1164
Structure of the Thermosphere—M. Nicolet. (*Planet. Space Sci.*, vol. 5, pp. 1–32; January, 1961.) The vertical distribution of density in the thermosphere as deduced from satellite observations is examined and it is shown that the varying scale-height gradient is due to a decrease in the molecular weight of the atmospheric constituents. The temperature of the nighttime isothermal atmosphere and the distribution of scale height in the sunlit atmosphere is related to the solar ultraviolet heating available during the day. 87 references.

551.510.535 1165
Origin of the D Region—E. C. Y. Inn. (*Planet. Space Sci.*, vol. 5, pp. 76–78; January, 1961.) The D region may result from the ionization, by Lyman alpha radiation, of O_2 molecules in an excited state. The mechanism for the excitation of the O_2 molecules is discussed.

551.510.535 1166
Anisotropy in Ionospheric Diffraction and its Effect on Drift Measurement—R. B. Banerji. [*Proc. Phys. Soc.* (*London*), vol. 76, pp. 959–968; December 1, 1960.] A method of analysis is developed in which five statistical parameters and a few simple formulas are used, and which takes into account the effects of random motion as well as anisotropy.

551.510.535 1167
Lunar Tide in the F_2 Layer of the Ionosphere near the Geomagnetic Equator—R. G. Rastogi. (*Nature*, vol. 189, pp. 214–215; January 21, 1961.) Lunar variations of $f\delta$ F_2 at Leopoldville during the period November, 1953–October, 1956 are analyzed and compared with those for other low-latitude stations. Results suggest that stations having the same magnetic latitude rather than the same geomagnetic latitude experience similar variations of $f\delta$ F_2. See also 866 of March (Bossolasco and Elena).

551.510.535 1168
Spread-F and Multiple Scattering in the Ionosphere—D. S. Bugnolo. (*J. Geophys. Res.*, vol. 65, pp. 3925–3929; December, 1960.) The mechanism of spread-F is examined statistically. A model is derived, based on the assumption of Gallet turbulence in the underside of the F layer under nighttime conditions, and is applied in detail to a typical example of arctic spread-F.

551.510.535:523.165 1169
Ionospheric Disturbances at the Time of Cosmic-Ray Increases—C. Collins and D. H. Jelly. (*Nature*, vol. 189, pp. 128–129; January 14, 1961.) Ionospheric and magnetic conditions at the time of a cosmic-ray increase on January 21, 1957 are analyzed and compared with those at the time of an increase on December 4, 1957. Although the cosmic-ray increases were in some respects analogous, they were associated with quite different ionospheric conditions.

551.510.535:551.594.6 1170
Spread-F and the Latitude Variation of Occurrence of Whistlers—D. G. Singleton. (*Nature*, vol. 189, pp. 215–216; January 21, 1961.) An explanation of the latitude variation of whistler activity based on propagation within field-aligned ducts [2753 of 1960 (Smith, *et al.*)] is consistent with current ideas for interpreting spread-F echoes [see *e.g.*, 1737 of 1958 (Briggs)].

551.510.535:523.75 1171
Enhancement of Ionization in the E Layer due to Solar Flares during the International

Geophysical Year—G. H. Bazzard. (*Nature*, vol. 189, pp. 47–48; January 7, 1961.) The value of f_0E observed at Slough at the hour preceding the commencement of each flare of importance 2^+ and 3, and at the two succeeding hours, has been examined in relation to normal values. Results show that during the interval of 25–120 minutes after the commencement of a flare there is a mean increase in the intensity of E-layer ionizing radiation which is at least 22 per cent for class 3 flares and 8 per cent for class 2^+ flares.

551.510.535:621.391.812.63 1172
On the Generalization of the Appleton-Hartree Magneto-ionic Formulas—Sen and Wyller. (See 1296.)

551.510.535(98):621.391.812.631 1173
The Polar Radio Black-Out of the Ionosphere of 7 July, 1958—J. H. Chapman. (*Canad. J. Phys.*, vol. 38, pp. 1195–1212; September, 1960.) Ionospheric data are used to study the polar-cap absorption effects occurring in both hemispheres after the 3^+ flare on July 7. It is suggested that electrons first precipitated to the highest latitudes and then protons and helium ions to lower latitudes.

551.594.5:551.510.535 1174
A Theory of Ionospheric Currents Associated with Aurorae: Parts 1 & 2—J. T. Weaver and R. Skinner. (*Canad. J. Phys.*, vol. 38, pp. 1089–1113; August, 1960.) The current distributions induced by dynamo action in a sheet of partially ionized gas moving with uniform velocity across a perpendicular magnetic field are derived for the case when the gas contains an elliptical region of greater ionization density. On the basis of the theory, the average ionospheric current flowing along a long homogeneous arc is calculated and some results predicted by the theory are compared with observational evidence.

551.594.5:621.396.96 1175
Studies of Auroral Echoes: Part 1—L. Harang and J. Tröim. (*Planet. Space Sci.*, vol. 5, pp. 33–45; January, 1961.) The results of 40-Mc radar observations at Kjeller (60°N) and Tromsö (70°N) are discussed. The velocity and direction of the drift of auroral ionization was measured by the technique used at Jodrell Bank by Bullough and Kaiser (2622 of 1955), in which the time variation of echoes observed on antennas directed NW and NE is recorded. An interferometer was used to measure fine structure, and a vertically sliding antenna to measure the angle of elevation of the arriving echoes. The observed drift and structure at Kjeller was similar to that at Jodrell Bank, but was more irregular at Tromsö.

551.594.5:621.396.96 1176
Radar Echoes from the Aurora at 1300 Mc/s—E. Eastwood, G. A. Isted, and J. D. Bell. (*Nature*, vol. 189, pp. 115–117; January 14, 1961.) Experimental results show that the reflecting layer has a mean height of 90 km and a thickness of 20 km. The results are discussed in relation to reflection theory. In general, an auroral disturbance of sufficient intensity to be observed at 1300 Mc will only occur after a very vigorous solar upheaval in association with extreme magnetic storms.

551.594.6 1177
Recent Results in the Investigation of the Relation between Lightning Discharges and Whistlers—H. Norinder and E. Knudsen. (*Planet. Space Sci.*, vol. 5, pp. 46–49; January, 1961.) Harmonic analysis shows that lightning discharges which produce whistlers have a pronounced energy peak near 5 kc. Multiple whistlers are caused by multiple discharges through the same lightning channel. See also 2383 of 1960.

LOCATION AND AIDS TO NAVIGATION

621.396.933.23 1178
The Problem of Improving the British Instrument Landing System Localizer for Automatic Landing—A. N. Beresford and J. D. Asteraki. (*Proc. IEE*, vol. 108, pt. B, pp. 59–64; January, 1961.) A discussion of the factors involved in obtaining an improvement in accuracy in the B.I.L.S. localizer is given.

621.396.96 1179
Two Statistical Models for Radar Terrain Return—L. M. Spetner and I. Katz. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 242–246; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1511; August, 1960.)

621.396.96 1180
Radar Terrain Return Measured at Near-Vertical Incidence—A. R. Edison, R. K. Moore, and B. D. Warner. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 246–254; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1511; August, 1960.) See also 1457 of 1957 (Moore and Williams).

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215 1181
Mechanism of Photoconductivity in Microcrystalline Powders—R. H. Bube. (*J. Appl. Phys.*, vol. 31, pp. 2239–2254; December, 1960.) Measurements on a variety of CdS powder cells show the four different conduction regions in the I/V characteristic associated with two barrier breakdown processes. An explanation is proposed and a model constructed.

535.215 1182
Field-Effect Modulation of Photoconductance in a Quasi-intrinsic Semiconductor—R. R. Bockemuehl. (*J. Appl. Phys.*, vol. 31, pp. 2255–2259; December, 1960.) An analysis of electron-pair generation in a depletion layer explains the terminal characteristics of a photoconductive field-effect transistor and the optical and frequency response of field-effect modulation on the photoconductance of CdS. It also gives a method for evaluating carrier properties in a quasi-intrinsic semiconductor.

535.215:539.23 1183
The Photoeffect of Thin Metal Films—F. Baumann. (*Z. Phys.*, vol. 158, pp. 607–622; April 12, 1960.) The photoelectric emission of thin layers of Pb, Sn, Bi and Sn-Cu, prepared by quenching at low temperatures, was measured immediately after condensation and after difference annealing processes. The effects of annealing on photoelectric sensitivity and threshold position are discussed.

535.215:546.31'86 1184
Photoemission and Related Properties of the Alkali Antimonides—W. E. Spicer. (*J. Appl. Phys.*, vol. 31, pp. 2077–2084; December, 1960.)

535.215:546.32'33'86 1185
Phases in the Photoelectric Sodium-Potassium-Antimony System—W. H. McCarroll. (*J. Phys. Chem. Solids*, vol. 16, pp. 30–36; November, 1960.)

535.215:546.47'221 1186
The Piezo-optic and Electro-optic Constants of Zinc-blende—R. Bechmann. (*J. Phys. Chem. Solids*, vol. 16, pp. 100–101; November, 1960.)

535.215:[546.47'231+546.47'241 1187
The Optical and Photoelectric Properties of Zinc Selenide and Telluride—G. A. Zholkevich. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1115–1117; June, 1960.) Measurements of spectral distribution of the photocurrent made on more than 100 samples showed that ZnSe crystals have a principal maximum of photoconductivity at about 460 mμ in the intrinsic absorption region; ZnTe crystals did not show a maximum but a plateau in the region of 520 mμ.

535.215:[546.47'241+546.48'241 1188
Electrical Conductivity and Photoelectric Properties of Cadmium and Zinc Telluride Layers—P. P. Konorov and I. B. Shevchenko. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1134–1140; June, 1960.) Vacuum-deposited CdTe films 1 μ thick show high photosensitivity in the visible and near-infrared part of the spectrum, and could be used as photoresistors. The mechanism of sensitization of these films is discussed. ZnTe did not display appreciable photosensitivity.

535.215:546.48'221 1189
Fluctuations, Neutral Transport and the Photoconductive Process in CdS—R. J. Robinson and J. J. Brophy. (*Physica*, vol. 26, pp. 440–442; June, 1960.) The noise spectra of single crystals of CdS lightly doped with CuCl have been studied under both uniform and masked (photodiffusion) illumination, using 5200-Å radiation. Results obtained under photodiffusion conditions show an additional noise process which appears to have a relaxation time of 500 μsec.

535.215:546.48'221 1190
The Influence of Gaseous Atmospheres on the Spectral Distribution of Photoconductivity in CdS Single Crystals—H. Berger, K. W. Böer, and E. H. Weber. (*Z. Phys.*, vol. 158, pp. 501–510; April 12, 1960.) Investigations cover the influence of O_2, N_2, H_2, H_2O, CO_2 on undoped CdS single crystals; measurements were made near the absorption edge at room temperature.

535.215:546.48'221 1191
Distribution of Electrons in Cadmium Sulphide Crystals—G. P. Mohanty and L. V. Azaroff. (*Phys. Rev.*, vol. 120, pp. 1224–1225; November 15, 1960.)

535.215:546.48'221 1192
Some Characteristics of Thermally Induced Conductivity in Single Crystals of CdS—A. P. Trofimenko, G. A. Fedorus, and A. K. Razmadze. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1141–1147; June, 1960.) Investigation showed the capture cross section for electrons in CdS crystals with excess S to be dependent on temperature. An experimental test was carried out to check the extent to which traps were filled when the thermally-induced conductivity was a maximum.

535.215:546.48'221 1193
Photoconductivity Mechanism in CdS-Type Single Crystals—M. K. Sheĭnkman. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1155–1159; June, 1960.) An analysis of experimental data based on the exciton hypothesis is given.

535.215:546.48'221 1194
Investigation of the Kinetics of Infrared Impurity Photoconductivity in CdS, Induced by Preliminary Illumination—E. N. Arkad'eva, L. G. Paritskiĭ, and S. M. Ryvkin. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1160–1168; June, 1960.) Investigation is made of the photoconductivity of CdS crystals at 77°K in the infrared region from 2 to 4 μ produced by pre-excitation due to illumination at a wavelength corresponding to the intrinsic absorption. A model for the infrared photoconductivity is discussed, based on the transition of electrons to the conduction band from trapping levels which are filled with electrons resulting from the preliminary illumination.

535.37:546.48'221 1195
Polarization of the Edge Emission in CdS—R. J. Collins and J. J. Hopfield. (*Phys. Rev.*, vol. 120, pp. 840–842; November, 1960.)

535.376 1196
(Zn, Hg)S and (Zn, Cd, Hg)S Electroluminscent Phosphors—A. Wachtel. (*J. Electrochem. Soc.*, vol. 107, pp. 682–688; August, 1960.) Procedures for the preparation of solid solutions of cubic (Zn, Hg)S and (Zn, Cd, Hg)S by firing in sealed silica tubes are described. With suitable additions of Cu as activator and of halides Ga or In as coactivator, photoluminescence and electroluminescence have been obtained. The electroluminescence at the red end of the spectrum consists of two emission bands which do not appear to be analogous to the blue and greem emission bands of Cu, Cl and ZnS.

535.376:546.47'221 1197
Electrical Properties of Electroluminescent ZnS Powders—R. Goffaux. (*J. Phys. Radium*, vol. 20, suppl. to no. 4, *Phys. Appl.*, pp. 18A–22A; April, 1959.) The properties of ZnS powder have been discussed and interpreted with the aid of varistor theory. The layer impedance may be represented by a variable resistance R_p in parallel with a variable capacitance C_p. An interpretation of the variation of R_p and C_p with voltage has been developed from previous work on varistors subjected to an alternating voltage (3135 of 1958); experimental results are in good agreement with this interpretation.

535.376:546.681'17 1198
Luminescence and Photoconductivity Properties of Doped GaN—H. G. Grimmeiss, R. Groth, and J. Maak. (*Z. Naturforsch.*, vol. 15a, pp. 799–806; September, 1960.) Similarities with the characteristics of ZnS-type phosphors were found.

537.226:621.396.677.85 1199
Some Theoretical Investigations on Metal-Disc Delay Dielectrics—S. K. Chatterjee and C. Dhanalkashmi. (*J. Inst. Telecommun. Engrs., India*, vol. 6, pp. 149–152; June, 1960.) The phase-shift suffered by a plane EM wave passing through an array of metal disks is calculated. The dielectric constant of the medium is also found.

537.226.2 1200
Measurement of Complex Dielectric Constant of Rochelle Salt at a Frequency of 10 Gc/s as a Function of Temperature and Electrical Bias Voltage—W. Jäckle. (*Z. angew. Phys.*, vol. 12, pp. 148–155; April, 1960.)

537.227 1201
Electro-optical Kerr Effect and Polarization Reversal in Deuterium-Doped Rochelle Salt—H. H. Wieder and D. A. Collins. (*Phys. Rev.*, vol. 120, pp. 725–730; November 1, 1960.) Polarization reversal as a function of nucleation and growth of domains in ferroelectric deuterium-doped Rochelle salt was investigated by means of the Kerr effect.

537.227:537.5 1202
High-Temperature Discharges in Ferroelectric Ceramics—J. W. Northrip. (*J. Appl. Phys.*, vol. 31, pp. 2293–2296; December, 1960.) An experimental investigation of the properties of a temperature-dependent current discharge seen in disks of $BaTiO_3$ and $Pb(Zr, Ti)O_3$ is described. This discharge, which is largest between 300 and 600°C, and falls off exponentially with time, depends primarily on the electrode material and its application. An explanation is proposed in terms of solid-state chemical activity.

537.227:546.431'824-31 1203
Aging of Barium Titanate Single Crystals—A. Misarova. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1276–1282; June, 1960.) A description of measurements of the changes of permittivity tan δ, the hysteresis loop and the conductivity of $BaTiO_3$ single crystal during aging, with and without a constant electric field, is given. A hypothesis analogous to that of Müser (3008 of 1959) is proposed to explain the aging process.

537.311.31:538.63 1204
The Relation between the Temperature Dependence of Electrical Resistance at Low Temperatures and the Galvanomagnetic Effect in Strong Magnetic Fields—O. S. Galkina and L. A. Chernikova. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 3–6; January, 1960.) Results of measurements on Ni-Cu alloys support the assumption that the change in electrical resistance is due to the scattering of conduction electrons by the inhomogeneities in lattice magnetization.

537.311.33 1205
Determination of Carrier Mobility and Density in the Surface Layer of a Semiconductor—V. K. Subashiev and S. A. Poltinnikov. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1169–1177; June, 1960.) Two methods are described for determining the surface carrier density and mobility of a semiconductor layer formed by impurity diffusion. The methods are based on Hall-effect and conductivity data and the relation between carrier mobility and density. Results of diffusion-layer measurements on Si photocells are given.

537.311.33 1206
Optical Absorption and Recombination Radiation in Semiconductors due to Transitions between Hydrogen-Like Acceptor Impurity Levels and the Conduction Band—D. M. Eagles. (*J. Phys. Chem. Solids*, vol. 16, pp. 76–83; November, 1960.) A simplified theory is given of the expected form of the absorption and radiation spectra. Necessary modifications are discussed and a comparison made with experimental observations.

537.311.33 1207
Impurity Conduction at Low Concentrations—A. Miller and E. Abrahams. (*Phys. Rev.*, vol. 120, pp. 745–755; November 1, 1960.) The conductivity of an n-type semiconductor has been calculated in the region of low temperature and low impurity conduction using the model of phonon-induced "hopping" of electrons from donor site to donor site. The results are in satisfactory agreement with experiment.

537.311.33 1208
Filled and Empty Dangling Bonds in III–V Compounds—D. B. Holt. (*J. Appl. Phys.*, vol. 31, pp. 2231–2232; December, 1960.) An introduction of the idea of resonance to a model of dangling bonds which is free from the objections of the previous model [1693 of 1960 (Gatos, *et al.*)] is given.

537.311.33:538.569.4 1209
Properties of Semiconductors with an Extremum Loop: Part 1—Cyclotron and Combinational Resonance in a Magnetic Field Perpendicular to the Plane of the Loop—É. I. Rashba. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1224–1238; June, 1960.) Investigation of the absorption of radio waves in semiconductors having a specific band structure in which an extremum is not reached at isolated points, but over a curve in **k**-space is discussed. As a result of spin-orbit coupling, transitions involving a change in spin due to the Lorentz force are significantly strong.

537.311.33:546.23 1210
The Hole Mobility in Selenium—W. E. Spear. [*Proc. Phys. Soc.* (*London*), vol. 76, pp. 826–832; December 1, 1960.] At 20°C the effective hole mobility of vitreous Se lies between 0.13 and 0.14 cm²/v-sec and its temperature dependence shows it to be controlled by a level of acceptor states lying 0.14 ev above the valence band. The above values are similar to those for single crystals of hexagonal Se, suggesting a common controlling mechanism.

537.311.33:546.24 1211
Some Effects Occurring in Dislocated Tellurium—J. S. Blakemore, J. W. Schultz, and K. C. Nomura. (*J. Appl. Phys.*, vol. 31, pp. 2226–2231; December, 1960.)

537.311.33:546.26-1:538.63 1212
de Haas-van Alphen Effect in Graphite between 3 and 85 Kilogauss—W. J. Spry and P. M. Scherer. (*Phys. Rev.*, vol. 120, pp. 826–829; November 1, 1960.)

537.311.33:546.28 1213
Visible Light Emission and Microplasma Phenomena in Silicon p-n Junction: Part 2—Classification of Weak Spots in Diffused p-n Junctions—M. Kikuchi. (*J. Phys. Soc. Japan*, vol. 15, pp. 1822–1831; October, 1960.) Two groups of weak spots exist; one type shows microplasma current pulses at certain bias voltages and the other type contributes a "soft" component to the reverse current. Typical experimental results are given. Part 1: 4294 of 1960 (Kikuchi and Tachikawa).

537.311.33:546.28 1214
The Identification of Precipitate Particles in Single Crystals of Silicon by Reflection Electron Diffraction—R. C. Newman. [*Proc. Phys. Soc.* (*London*), vol. 76, pp. 993 996; December 1, 1960.]

537.311.33:546.28 1215
Spin-Lattice Relaxation of Donor Electrons in Silicon—J. Kondo. (*Prog. Theoret. Phys.*, vol. 24, pp. 161–170; July, 1960.) The relaxation rate is calculated using an adiabatic approximation.

537.311.33:546.28 1216
Interstitial versus Substitutional Oxygen in Silicon—W. L. Bond and W. Kaiser. (*J. Phys. Chem. Solids*, vol. 16, pp. 44-45; November, 1960.) "Lattice constant and density studies on oxygen free and oxygen doped silicon single crystals suggest that oxygen occupies an interstitial lattice site."

537.311.33:546.28 1217
Gold-Induced Climb of Dislocations in Silicon—W. C. Dash. (*J. Appl. Phys.*, vol. 31, pp. 2275–2283; December, 1960.

537.311.33:546.281'26 1218
Nonlinear Properties and Ionic Migration Carbide—R. Goffaux. (*Rev. gén élect.*, vol. 69, pp. 331–338; June, 1960.) Experimental results are interpreted by a hypothesis in which mobile ions are held in intergrain boundaries by the field of surface-adsorbed charges and move under the influence of temperature or electric field. See also 3135 of 1958.

537.311.33:546.289 1219
Hot Carriers in Germanium—B. V. Rollin and J. M. Rowell. [*Proc. Phys. Soc.* (*London*), vol. 76, pp. 1001–1002; December 1, 1960.] Experiments on the variation of Hall mobility with lattice temperature and incident radiation are described.

537.311.33:546.289 1220
Investigations of Wetting and Alloying on Germanium—N. Michelitsch. (*Z. angew. Phys.*, vol. 12, pp. 180–184; April, 1960.) The difficulties of wetting and alloying In and In alloys with n-type Ge, and the means of overcoming them, are investigated.

537.311.33:546.289 1221
Evaluation and Control of Diffused Impurity Layers in Germanium—H. S. Veloric and W. J. Greig. (*RCA Rev.*, vol. 21, pp. 437–456; September, 1960.) A new solid-phase diffusion process with powdered Ge alloy of known impurity concentration is evaluated. Experiments show that the surface concentration of added impurities can be estimated from the impurity concentration in the source Ge powder.

537.311.33:546.289 1222
Thermal Conductivity of Germanium from 3°K to 1020°K—G. A. Slack and C. Glassbrenner. (*Phys. Rev.*, vol. 120, pp. 782–789; November 1, 1960.)

537.311.33:546.289 1223
Low-Temperature Transport in 'Split *p*-Germanium'—S. H. Koenig and J. J. Hall. (*Phys. Rev. Lett.*, vol. 5, pp. 550–553; December 15, 1960.) Measurements of conductivity and Hall mobility in Ge crystals, at temperature below 10°K and under large shear strains, give information concerning deformation potentials and velocity dependence of the recombination cross section.

537.311.33:546.289 1224
Valley-Orbit Splitting of Antimony in Germanium—H. Fritzsche. (*Phys. Rev.*, vol. 120, pp. 1120–1124; November 15, 1960.) The electrical conductivity of Sb-doped Ge has been measured in single crystals subjected to stress. The valley-orbit splitting was thereby deduced to be about 0.57 mev.

537.311.33:546.289 1225
Infrared Investigation of the Acceptor Levels formed by Copper in Germanium—D. L. Greenaway. [*Proc. Phys. Soc.* (*London*), vol. 76, pp. 900–908; December 1, 1960.] Measurements between 1 and 34 μ indicate that the three acceptor levels introduced can be identified by their effect on the low-temperature absorption spectra of suitably doped samples.

537.311.33:546.289 1226
Note on Semiconductor Statistics—S. Teitler and R. F. Wallis. (*J. Phys. Chem. Solids*, vol. 16, pp. 71–75; November, 1960.) The Guggenheim method using the grand partition function is applied to multilevel impurities in Ge.

537.311.33:546.289 1227
Electrical Properties of Heavily Doped *n*-Type Germanium—Y. Furukawa. (*J. Phys. Soc. Japan*, vol. 15, pp. 1903–1904; October, 1960.) The electron concentration in four crystals doped with both As and Sb has been compared with resistivity. *I/V* characteristics of Esaki diodes made with the material have been studied.

537.311.33:546.289 1228
The Diffusion of Hydrogen in Single-Crystal Germanium—R. C. Frank and J. E. Thomas, Jr. (*J. Phys. Chem. Solids*, vol. 16, pp. 144–151; November, 1960.) Permeation rates and diffusion coefficients were measured between 800 and 910°C.

537.311.33:546.289–31 1229
Donor Equilibria in the Germanium-Oxygen System—C. S. Fuller W. Kaiser and C. D. Thurmond. (*J. Phys. Chem. Solids* vol. 16, pp. 161–163; November, 1960.) Experimental results show agreement with theory revised to include the effect of intrinsic carrier concentration.

537.311.33:546.47'86 1230
Anisotropy of Certain Electrical Properties of Single Crystals of Zinc Antimonide—M. V. Kot and I. V. Kretsu. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1250–1255; June, 1960.) Methods of preparation for single crystals of ZnSb are outlined and measurements of their electrical properties are reported. The crystals show *p*-type conductivity and anisotropy in their electrical properties.

537.311.33:546.48–31:539.23 1231
The Relations between the Optical Absorption and Dispersion and the Conductivity Mechanism of Thin Cadmium Oxide Films—H. Finkenrath. (*Z. Phys.*, vol. 158, pp. 511–532; April 12, 1960.)

537.311.33:546.48'86 1232
Anisotropy of Electrical Properties of Single Crystals of Cadmium Antimonide—I. K. Andronik and M. V. Kot. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1128–1133; June, 1960.) Methods of preparing single crystals of CdSb are described and the temperature dependence of electrical conductivity, thermal EMF and Hall effect are examined. Crystals have anisotropic electrical properties and the width of the forbidden band is about 0.57 ev.

537.311.33:546.681'19 1233
Space-Charge Currents in Gallium Arsenide—J. W. Allen and R. J. Cherry. (*Nature*, vol. 189, pp. 297–298; January 28, 1961.) Experimental results are presented confirming the correctness of a model proposed earlier by Allen (3575 of 1960) to account for the properties of semi-insulating GaAs.

537.311.33:546.681'4'241 1234
Effects of Solid Solution of Ga_2Te_3 with $A^{II}B^{VI}$ Tellurides—J. C. Woolley and B. Ray. (*J. Phys. Chem. Solids*, vol. 16, pp. 102–106; November, 1960.)

537.311.33:546.681'86 1235
Thermomagnetic Properties of Gallium Antimonide—D. Kh. Amirkhanova. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1125–1127; June, 1960.) An investigation of the Nernst-Ettingshausen effects on five *p*-type GaSb specimens in the temperature range 100–900°K is made.

537.311.33:546.682'19 1236
Voigt Effect in Semiconductors—S. Teitler and E. D. Palik. (*Phys. Rev. Lett.*, vol. 5, pp. 546–548; December 15, 1960.) The effective electron mass ratio for *n*-type InAs is determined from the magnetic double refraction of 15-μ infrared radiation in fields up to 70 kg.

537.311.33:[546.682'86+546.682'19'18 1237
Hall Constant and Electron Mobility of InSb, InAs and $In(As_{0.8}P_{0.2})$ in Strong Magnetic Fields—E. Braunersreuther, F. Kuhrt, and H. J. Lippmann. (*Z. Naturforsch.*, vol. 15a, 795–799; September, 1960.) Measurements were made at field strengths up to 180 kg. The Hall constants of InAs and $In(As_{0.8}P_{0.2})$ were found to be independent of the field; in InSb the Hall constant dropped with increasing field strength (about 15 per cent at 170 kg).

537.311.33:546.682'86 1238
Lattice Absorption Bands in Indium Antimonide—S. J. Fray, F. A. Johnson and R. H. Jones. [*Proc. Phys. Soc.* (*London*), vol. 76, pp. 939–948; December 1, 1960.]

537.311.33:546.682'86 1239
Effect of Landau Levels upon Tunnel Currents in Indium Antimonide—A. G. Chynoweth, R. A. Logan and P. A. Wolff. (*Phys. Rev. Lett.*, vol. 5, pp. 548–550; December 15, 1960.) The tunnel current is observed to decrease in an oscillatory manner with an increase of magnetic field. This effect is attributed to fluctuations in level density at the Fermi surface.

537.311.33:546.682'86 1240
Magnetoresistance of High-Purity InSb in the Quantum Limit—R. J. Sladek. (*J. Phys. Chem. Solids*, vol. 16, pp. 1–9; November, 1960.) Measurements between 111°K and 50°K with magnetic field strengths up to 28 kg show qualitative agreement with theory for the quantum limit in the case of piezoelectric scattering and classical statistics.

537.311.33:546.682'86 1241
The Galvanomagnetic Properties of InSb Single Crystals with Te Doping—H. Rupprecht, R. Weber and H. Weiss. (*Z. Naturforsch*, vol. 15a, pp. 783–794; September, 1960.) Measurements were made on material with Te concentration $>10^{16}/cm^3$. The investigations cover the change of resistivity in a magnetic field as a function of doping, crystal orientation, magnetic flux density, and the angle between magnetic flux and current, at 295, 78 and 4.2°K.

537.311.33:546.682'86'241 1242
Some Electrical and Optical Properties of $InSb-In_2Te_3$ Alloys—J. C. Woolley, C. M. Gillett and J. A. Evans. (*J. Phys. Chem. Solids*, vol. 16, pp. 138–143; November, 1960.) The range of solid solution at the InSb end of the alloy system has been determined by X-ray methods to be about 15 mol per cent In_2Te_3 and properties in this range have been investigated.

537.311.33:546.873'241 1243
Theory of Piezoresistance in Bi_2Te_3—M. I. Klinger. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1353–1356; June, 1960.)

537.312.62 1244
Magnetic Moment of Transition Metal Atoms in Dilute Solution and their Effect on Superconducting Transition Temperature—B. T. Matthias, M. Peter, H. J. Williams, A. M. Clogston, E. Corenzwit, and R. C. Sherwood. (*Phys. Rev. Lett.*, vol. 5, pp. 542–544; December 15, 1960.) $Mo_{0.8}Re_{0.2}$ alloys show strong lowering of the transition temperature with increasing Fe concentration. The magnetic moment per Fe atom, in Nb-Mo solid solutions increases sharply with increase in Mo concentration.

537.323 1245
Thermoelectric Properties of Alloys of the Pseudo-binary System Sb_2Te_3-Bi_2Te_3—G. V. Kokosh and S. S. Sinani. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1118–1124; June, 1960.) An investigation is made of the effect of impurities on solid solutions of semiconductors with disturbed stoichiometry. The effect of firing on the electrical conductivity is examined and the properties of alloys of different purity are compared.

537.327:534.1 1246
Dependence of Mechanical *Q* and Young's Modulus of Ferroelectric Ceramics on Stress Amplitude—R. Gerson. (*J. Acoust. Soc. Am.*, vol. 32, pp. 1297–1301; October, 1960.) Measurements on $BaTiO_3$ and $Pb(Ti, Zr)O_3$ compositions are reported. The mechanical *Q* decreased to about one third of its low-signal value at a stress of 1500 psi. No further decrease occurred up to 4000 psi. Young's modulus decreased by about 5 per cent in materials used in power transducers, and by about 25 per cent in compositions having low coercive force.

538.22:537.311.33 1247
Electrical Conduction in Antiferromagnetics—G. L. Sewell. [*Proc. Phys. Soc.* (*London*), vol. 76, pp. 985–987; December 1, 1960.] The mechanism of phonon-induced electron jump is used to explain the anomalous electrical properties of NiO and α-Fe_2O_3.

538.22:546.824–31 1248
Magnetic Susceptibility of Tetragonal Titanium Dioxide—F. E. Senftle, T. Pankey, and F. A. Grant. (*Phys. Rev.*, vol. 120, pp. 820–825; November 1, 1960.) Measured values for the magnetic susceptibility are given as 0.067×10^{-6} emu/g for a single crystal of high-purity rutile and less than 0.02×10^{-6} emu/g for anatase powder.

538.22:546.824'881–31 1249
Magnetic Susceptibility of Solid Solutions of Vanadium Dioxide in Titanium Dioxide—S. M. Ariya and G. Grossmann. (*Fiz. Tverdogo Tela*, vol. 2, pp. 1283–1286; June, 1960.) A report of measurements on VO_2-TiO_2 solid solutions prepared by prolonged annealing at 800°C is given.

538.221 1250
Magnetic Viscosity due to Solute Atom Pairs: Part 1—Theory of the Effect—G. Biorci, A. Ferro, and G. Montalenti. (*J. Appl. Phys.*, vol. 31, pp. 2121–2125; December, 1960.) The additional effect of solute atom pairs in ferromagnetic alloys is explained.

538.221 1251
Exchange Anisotropy in Cobalt-Manganese Alloys—J. S. Kouvel. (*J. Phys. Chem. Solids*, vol. 16, pp. 107–114; November, 1960.)

538.221 1252
Time Decrease of Permeability of Substitutional Ferromagnetic Alloys—C. Kuroda. (*J. Phys. Soc. Japan*, vol. 15, p. 1898; October, 1960.) Results for Ni-Co alloys are given.

538.221 1253
Hall Effect and Resistivity of Ni-Pd Alloys—J. A. Dreesen and E. M. Pugh. (*Phys. Rev.*, vol. 120, pp. 1218–1233; November 15, 1960.)

538.221:537.311.31 1254
The Electrical-Resistance Maximum for Ferromagnetics at the Curie Points at Low Temperatures—E. I. Kondorskiĭ, O. S. Galkina, and L. A. Chernikova. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 646–648; February, 1960.) Experimental verification of the existence of a resistivity maximum in the region of magnetic saturation for Cu-Ni alloys with Curie point below 20°K is given.

538.221:538.632 1255
Temperature Dependence of the Hall Effect in Pure Ferromagnetics—N. V. Volkenshtein and G. V. Federov. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 64–67; January, 1960.) Results are given of measurements of the Hall effect in 99.998 per cent pure Fe, Ni and Co from room temperature down to 4.2°K. Present theories are inadequate for an explanation of the experimental data obtained.

538.221:621.317.441 1256
Homogeneous Magnetic Fields for Optical Investigations of Ferromagnetic Specimens—H. Murrmann and C. Schwink. (*Z. angew. Phys.*, vol. 12, pp. 155–157; 1960.) The design of solenoids producing uniform magnetic fields and provided with a gap for the observation of specimens being magnetized is discussed.

538.221:621.318.124 1257
The Effect of Cation Vacancies on the Magnetic Annealing of Co_x $Fe_{3-x}O_4$—T. Inoue, H. Mizuta, and S. Iida. (*J. Phys. Soc. Japan*, vol. 15, pp. 1899–1900; October, 1960.) The relaxation time is found to be inversely proportional to the density of cation vacancies in a certain range.

538.221:621.318.124 1258
Numerical Evaluation of Functions Occurring in a Study of Domain Configuration in Thin Layers of $BaFe_{12}O_{19}$—A. J. W. Duijvestijn and B. P. A. Boonstra. (*Philips Res. Repts.*, vol. 15, pp. 390–393; August, 1960.) Graphs are represented showing the variation of $(\delta_1+\delta_2)$ as a function of τ, and $N(\mu, \phi)$ as a function of u. See 2845 of 1960 (Kooy and Enz.)

538.221:621.318.132 1259
Contribution of Dislocations to the Initial Susceptibility of Magnetically Soft Material—G. Rieder. (*Z. Naturforsh.*, vol. 15a, pp. 746–748; August, 1960.) A note on the theory of magnetization reversal in weak fields with regard to the effect of dislocations on the deformation of Bloch walls is given. See 2467 of 1960.

538.221:621.318.134 1260
Contribution to the Experimental Investigation of the Properties of certain Ferrites in the 10-Gc/s Band—V. Cagan. [*Ann. Phys. (Paris)*, vol. 5, pp. 1301–1354; September/October, 1960.)] A detailed report of Faraday-effect and resonance measurements relating to the development of low-loss ferrites is given. 52 references.

538.221:621.318.134 1261
The Statistics of Superexchange Interaction and Ionic Distribution in Substituted Ferrimagnetic Rare-Earth Iron Garnets—S. Geller. (*J. Phys. Chem. Solids*, vol. 16, pp. 21–29; November, 1960.)

538.221:621.318.134 1262
Subsidiary Resonance in the Coincidence Region in Yttrium Iron Garnet—F. C. Rossol. (*J. Appl. Phys.*, vol. 31, pp. 2273–2275; December, 1960.) Measurements are made of the variation, with frequency, of h_{crit}, the threshold RF field for subsidiary resonance, for a single crystal sphere of Y-Fe garnet. The measured curve, flat from 2000–3300 Mc, is compared with one computed from Suhl's theory of subsidiary resonance at high-power levels (3052 of 1958).

538.221:621.318.134 1263
On Some Calcium-Iron-Oxygen Compounds—P. B. Braun and W. Kwestroo. (*Philips Res. Repts.*, vol. 15, pp. 394–397; August, 1960.) A description of the preparation and magnetic properties of the three new classes of compounds, stabilized by the addition of small amounts of a third component, is given.

MATHEMATICS

511+513]:621.396.74 1264
The Mathematical Principles Underlying the Theoretical Planning of Transmitter Networks—Fastert. (See 1310.)

517.564.4:537.56 1265
Cartesian Tensor Scalar Product and Spherical Harmonic Expressions in Boltzmann's Equation—T. W. Johnston. (*Phys. Rev.*, vol. 120, pp. 1103–1111; November 15, 1960.)

517.918 1266
The Equivalence of the Taylor-Cauchy and Laurent-Cauchy Transform Analysis with Conventional Methods—E. V. Bohn: A. A. Wolf. (Proc. IRE, vol. 49, pp. 358–361; January, 1961.) Comment is made on 2853 of 1960 (Ku, *et al.*) and the author's reply is given.

517.949.8:517.54 1267
Contribution to the Modified Z-Transform Theory—E. I. Jury. (*J. Franklin Inst.*, vol. 270, pp. 114–129; August, 1960.) The complex convolution integral for both the Z-transform and modified Z-transform are discussed in detail and considered in relation to pulsed systems.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74):538.569.4 1268
Hyperfine Splitting of Rubidium-87—L. Essen, E. G. Hope, and D. Sutcliffe. (*Nature*, vol. 189, p. 298; January 28, 1961.) A brief note on measurements of the atomic beam value for Rb-87 in a mixture with Cs-133 is given. The value corrected for asymmetry and magnetic field is 6 834 682 614.0±1 cps.

621.317.3:550.372 1269
Ground-Constant Measurements using a Section of Balanced Two-Wire Transmission Line—E. J. Kirkscether. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 307–312; May, 1960.) Abstract, Proc. IRE, vol. 48, p. 1512; August, 1960.)

621.317.34 1270
Errors due to Losses in Measurements on Balanced Elements—M. Soldi. (*Alta Frequenza*, vol. 29, pp. 30–58; February, 1960.) The errors which affect measurements on balanced circuits by means of baluns in the m-λ range are discussed. Accuracy can be improved by introducing corrections for the losses of the balun based on a modified equivalent circuit. Experimental results are given.

621.317.44 1271
Field Measurements by the Method of Harmonics: Methodical Investigations—J. Greiner. (*Nachrtech.*, vol. 10, pp. 123–126; March, 1960.) Differences in the characteristics of various types of magnetometer head are considered, and the results of comparative measurements are tabulated. See 2475 of 1960.

621.317.7:621.373.42.089.6 1272
The Concept of Equivalent Source E.M.F. and Equivalent Available Power in Signal-Generator Calibration—D. Woods. (*Proc. IEE*, vol. 108, pt. B, pp. 37–42; January, 1961.) A discussion of methods of standardizing the output level of signal generators, with special reference to the errors arising from mismatch, is given.

621.317.755:621.385.832.032.265 1273
Velocity Modulation and Related High-Frequency Deflection Errors of the Travelling-Wave Deflection System—Goldberg. (See 1378.)

621.317.789.029.63 1274
Standardization of Strong Electromagnetic Fields—J. A. Lane and J. E. Pearson. (*Nature*, vol. 189, pp. 49–50; January 7, 1961.) A brief description of a method for establishing standard flux densities under laboratory conditions at frequencies of the order of 1 Gc and input powers of 10–20 w is given. A TEM-type wave is transmitted between parallel brass plates 3.5 m long, 60 cm wide and 10 cm apart forming a strip-line, terminated in an absorbing load. Absolute values of the flux density in the strip-line may be determined from the voltage induced in a probe which has been calibrated by reference to a rectangular waveguide transmitting an H_{10} wave of known power. Preliminary results suggest an error in measurement of less than ±1 db.

621.317.794:621.372.822 1275
The Design and Performance of Transverse-Film Bolometers in Rectangular Waveguides—J. A. Lane and D. M. Evans. (*Proc. IEE*, vol. 108, pt. B, pp. 133–135; January, 1961.) Equations are derived which determine the design of film bolometers. Measurements on an improved tunable-type instrument at a wavelength of 3 cm indicate that power levels between 1 and 100 mw can be measured in terms of a dc calibration with an error of not more than ±2 per cent.

621.317.799:621.396.62.029.62 1276
Artificial Aerials for Measurements on Metre-Wave Reviewers—C. Egidi. (*Tech. Mitt. PTT*, vol. 38, pp. 66–101; March 1, 1960. In Italian and French.) Circuits, equations and tables of parameters are given for measurements with networks designed for coupling one or more signal generators to a receiver under various matching conditions.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.247.4:621.385.832.001.4 1277
C.R.T. Photomicrometer — A. Ciuciura. (*Mullard Tech. Commun.*, vol. 5, pp. 141–158; June, 1960.) A detailed description is given of an instrument which provides, on a pen-recorder chart, a light-intensity profile of a line on a cathode-ray tube raster.

621.362+621.56]:537.322 1278
Mathematical Theory of a Peltier-Effect Refrigerator and of a Thermoelectric Generator—E. B. Penrod and Cho Yen Ho. (*J. Phys. Radium*, vol. 21, suppl. to no. 7, *Phys. Appl.*, pp. 97A–112A; July, 1960.)

621.362:621.387 1279
Potential Distributions in a Low-Pressure Thermionic Converter—P. L. Auer. (*J. Appl. Phys.*, vol. 31, pp. 2096–2103; December, 1960.) A plane-diode model of a low-pressure Cs-filled thermionic converter is used for the analysis.

621.362:621.387 1280
Experimental Investigations of the Cesium Plasma Cell—W. Ranken, G. M. Grover, and E. W. Salmi. (*J. Appl. Phys.*, vol. 31, pp. 2140–2153; December, 1960.) Evaluation of some aspects of the performance of a Cs plasma cell with Ta emitter, and the effects of Cs vapor pressure, emitter temperature and emitter-collector separation is given.

621.38:551.46.018 1281
Electronic Techniques in Oceanography—M. J. Tucker. (*J. Brit. IRE*, vol. 20, pp. 921–931; December, 1960.) Underwater acoustics is discussed and a shipborne wave recorder and a Vibrotron FM pressure gauge are described.

621.383.8:77 1282
Image Converter Tube for High-Speed Photography—T. Nakamura and T. Kasai. (*Electronics*, vol. 33, pp. 76–78; December 9, 1960.) Shutter pulses 0.1 μsec wide repeated every 2 μsec combined with electron-beam deflection give a series of displaced images.

621.385.833 1283
Small-Angle Diffraction with Electron Beams—H. Mahl and W. Weitsch. (*Naturwiss.*, vol. 47, pp. 301–302; July, 1960.) A note on the application of an electron microscope for observations of lattice spacings up to about 1000 Å is given.

621.398 1284
Analogue Telemetry Equipment and Systems: Part 1—R. E. Young. (*Electronic Engrg.*, vol. 33, pp. 16–20; January, 1961.) An air-to-ground television installation and automatic radome test equipment are described.

PROPAGATION OF WAVES

621.391.812.62:621.397 1285
Weather Conditions during Television Reception from East Germany and Italy—Braam. (See 1320.)

621.391.812.62.029.65 1286
Propagation at 36000 Mc/s in the Los Angeles Basin—W. L. Flock, R. C. Mackey, and W. D. Hershberger. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 235–241; May, 1960. Abstract, Proc. IRE, vol. 48, p. 1511; August, 1960.)

621.391.812.63 1287
H. F. Propagation—its Present and Future Use for Communication Purposes—A. F. Wilkins. (*J. Brit. IRE*, vol. 20, pp. 939–951; December, 1960.) The history, current practices, and future trends in long-distance propagation are reviewed. Improved ionospheric predictions are possible by use of an index I_{F2} based on f_0F_2, by frequency-sweep, and by back-scatter techniques. Propagation effects on HF antenna design, and the use of earth satellites for VHF and UHF communication is mentioned. 28 references.

621.391.812.63 1288
Applying the Chordal-Hop Theory of Ionospheric Long-Range Propagation to Echo-Signal Delay—H. J. Albrecht. (Proc. IRE, vol. 49, pp. 356–357; January, 1961.) An analysis of published time delays of round-the-world echoes shows that the predominant mode of propagation is one in which the waves suffer no ground reflections but are reflected only by the ionosphere.

621.391.812.63 1289
On the Effect of a Magnetic Field on the Spectrum of Incoherent Scattering—T. Laaspere. (*J. Geophys. Res.*, vol. 65, pp. 3955–3959; December, 1960.) A theoretical analysis of scattering by electrons in the ionosphere neglecting es attraction between electrons and ions is given. The thermal velocity of the electrons causes them to gyrate about and drift along the magnetic lines. The gyration produces phase modulation of the incident wave giving a line spectrum. The drift velocity blurs the line spectrum if the scattering wave vector is not normal to the lines.

621.391.812.63 1290
Scattering of Radio Waves by an Ionized Gas in Thermal Equilibrium—J. A. Fejer. (*Canad. J. Phys.*, vol. 38, pp. 1114–1133; August, 1960.) Expressions for the frequency /power spectrum of waves scattered by density fluctuations are derived and used in the interpretation of results of ionospheric observations of this type of scattering in terms of electron density and temperature. A shortened revised report of this work was noted earlier (191*f* of January).

621.391.812.63 1291
Skip-Distance Ray Focusing in the Ionosphere—E. Golton. (*Nature*, vol. 189, pp. 48–49; January 7, 1961.) Observations of skip-distance focusing of 20-Mc signals from 1958δ are examined and found to be in general agreement with the related theoretical analysis of Bremmer.

o21.391.812.63 1292
Experimental Proof of Focusing at the Skip Distance by Back-Scatter Records—H. F. Bates, R. D. Hunsucker, and L. Owren: K. Bibl. (Proc. IRE, vol. 49, p. 369; January, 1961.) A comment on the distinction between least-time and skip-zone focusing (2880 of 1960) is made and the author's reply given.

621.391.812.63:531.76 1293
Method of Studying Travel Time Anomalies of High-Frequency Radio Waves—G. Lerfold and P. Scheibe. (*Rev. Sci. Instr.*, vol. 31, pp. 1309–1311; December, 1960.) A device is described for the automatic measurement of changes in travel time of signals propagated via the ionosphere from station WWV.

621.391.812.63:551.507.362 1294
Doppler Shifts and Faraday Rotation of Radio Signals in a Time-Varying Inhomogeneous Ionosphere: Part 1—Single Signal Case—J. M. Kelso. (*J. Geophys. Res.*, vol. 65, pp. 3909–3914; December, 1960.) The ionosphere is treated as quasi-isotropic; the ray path is obtained in a manner which is exact in an isotropic medium, but the refractive index is permitted to be a function of ray direction.

621.391.812.63:551.507.362.2 1295
A Reciprocity Theorem for Nonperiodic Fields—G. Goubau. (IRE Trans. on Antennas and Propagation, vol. AP-8, pp. 339–342; May, 1960.) Reciprocity in transmission between a satellite and a ground station is considered theoretically and is shown to be invalid due to both gyromagnetic effects in the ionosphere and to the fact that a moving source is involved.

621.391.812.63:551.510.535 1296
On the Generalization of the Appleton-Hartree Magneto-ionic Formulas—H. K. Sen and A. A. Wyller. (*J. Geophys. Res.*, vol. 65, pp. 3931–3950; December, 1960.) The Appleton-Hartree formulas are modified by assuming the collision frequency to be dependent on electron velocity. The elements of the generalized conductivity tensor are expressed as integrals which can be evaluated from existing tabulations in the case when the collision frequency is proportional to the square of the velocity. Numerical calculations are given for several cases of longitudinal and transverse propagation, and are compared with corresponding Appleton-Hartree results.

621.391.812.8 1297
Vertical and Oblique Incidence before the C.C.I.R. at Los Angeles, 1959—R. Gea Sacasa. [*Rev. Telecommun.* (*Madrid*), vol. 15, pp. 6–12; March, 1960.] A summary of statements emphasizing the superiority of predictions based on oblique-incidence measurements is given. A comparison is made between the results of Gea's method (*e.g.*, 1779 of 1960) and those obtained by the Japanese method [235 of 1956 (Miya and Kanaya)] for the circuit Tokyo–Geneva.

621.391.812.8:621.396.4 1298
The Variation of the Usable Daytime Frequencies in Short-Wave Long-Distance Traffic as an Effect of Solar Particle Radiation—G. Lange-Hesse. (*Arch. elekt. Übertragung*, vol. 14, pp. 115–120; March, 1960.) An analysis of observational data obtained at Lindau and Wingst to determine the statistical relation between the geomagnetic activity, used as a measure of solar particle radiation, and the relative deviation of the *M*3000 factor and $(M3000)F_2$ from their median values is given. A dependence on sunspot cycle and on seasons is found, and the significance of the results with regard to the prediction of disturbances is discussed.

RECEPTION

621.391.82:621.396.65 1299
Radio-Frequency Interference in Multichannel Telephony F.M. Radio Systems—R. Hamer. (*Proc. IEE*, vol. 108, pt. B, pp. 75–89; January, 1961.) The noise power ratio in the output telephony baseband due to a RF interfering signal is derived theoretically, and the results are compared with measured values.

621.396.62.029.62:621.317.799 1300
Artificial Aerials for Measurements on Metre-Wave Receivers—Egidi. (See 1276.)

621.396.666:621.396.4 1301
Multiple Diversity with Nonindependent Fading—J. N. Pierce. (Proc. IRE, vol. 49, pp. 363–364; January, 1961.) The analytical

method for determining multiple-diversity system performance, given by Pierce and Stein (1386 of 1960), is illustrated for the special case of binary data transmission by frequency-shift keying.

STATIONS AND COMMUNICATION SYSTEMS

621.376.2 1302
A General Theory and Oscilloscopic Study of Single Sideband Generation by Weaver's Method—S. Ramachandran. (*J. Inst. Telecommun. Engrs. India*, vol. 6, pp. 182–196; June, 1960.) Adjustments to Weaver's system (PROC. IRE, vol. 44, pp. 1703–1705; December, 1956) are best made by modulating with a sinusoid, and viewing a few cycles of RF on a repetitive timebase. A mathematical analysis of the pattern shapes is given.

621.391 1303
Information Theory—P. Holroyd and G. P. Jones. (*Electronic Tech.*, vol. 38, pp. 49–57; February, 1961.) A detailed introduction with practical examples is given.

621.395.44 1304
Design of the Terminal Equipments for the 12-Mc/s System on Coaxial Pairs—H. Iijima. (*Rev. Elect. Commun. Lab., Japan*, vol. 8, pp. 280-287; May/June, 1960.) The equipment for a 12-Mc wide-band system is outlined and may be added to existing 2.6/9.5-mm standard coaxial lines.

621.396.1 1305
The Economy of Radio Frequencies—S. Silleni. (*Alta Frequenza*, vol. 29, pp. 96–127; February, 1960.) The international regulations regarding frequency allocation are briefly reviewed. Ideal conditions for occupying the radio-frequency spectrum on the basis of various principles of "sharing" are considered, together with the practical limitation of such arrangements.

621.396.43:551.507.362.2 1306
A Transatlantic Communication Experiment via Echo I Satellite—H. Carru, R. Gendrin, and M. Reyssat. (*Nature*, vol. 189, pp. 268–271; January 28, 1961.) An analysis is given of the results of an experiment in which CW signals transmitted from Holmdel, N. J., on a frequency of 960.045 Mc were received at Issyles-Moulineaux, France, after reflection from the satellite 1960ι.

621.396.65 1307
Through-Switching Methods in Radio-Link Relay Stations—A. Egger. (*Rundfunktech. Mitt.*, vol. 4, pp. 80–84; April, 1960.) Block diagrams of switching systems for FM links are discussed and the effects on transmission quality and operation of the radio link are considered.

621.396.65 1308
Radio Links at Centimetre Wavelengths—F. Klima and R. Tuhl. (*Radio u. Fernsehen*, vol. 9, pp. 141–143; March, 1960.) Details are given of Czech equipment for mobile and fixed installations, mainly for use in television links in the 3- and 6-cm bands.

621.396.65:621.376.54 1309
The Multiplexing of Radio Link Systems with Pulse Width Modulation by Frequency Modulation of the R.F. Carrier—R. Ebermann. (*Nachrtech.*, vol. 10, pp. 95–103; March, 1960.) Block diagrams of transmitter and receiver using pulse width modulation with simultaneous frequency modulation of the carrier are given. The effect of FM on pulse shape and transmission characteristics is considered.

621.396.74:[511+513 1310
The Mathematical Principles Underlying the Theoretical Planning of Transmitter Networks—H. W. Fastert. (*Rundfunktech. Mitt.*, vol. 4, pp. 48–56; April, 1960.) Mathematical theory underlying the method given in 717 of February (Eden, *et al.*) is discussed. For the English version see *E.B.U. Rev.*, no. 60A, pp. 60–69; April, 1960.

621.396.97:534.76 1311
A Compatible Stereophonic System for the A.M. Broadcast Band—J. Avins, L. A. Freedman, F. R. Holt, J. H. O'Connell, J. O. Preisig, and R. N. Rhodes. (*RCA Rev.*, vol. 21, pp. 299–359; September, 1960.) Various possible systems are discussed. An AM-FM system using a pre-emphasized (left − right) signal to modulate the carrier frequency, with the resultant signal amplitude modulated by the (left + right) signal, was found to be superior.

SUBSIDIARY APPARATUS

621–526:621.313.2 1312
Speed Control of D.C. Motors—M. J. Miller and G. V. Buckley. (*Electronic Tech.*, vol. 38, pp. 63–67; February, 1961.) Two designs of a continuously variable type of servo system are described: the first employs a phase-lock feedback system and the second a simple velocity feedback system.

621–526:681.142:621.317.727 1313
Multitapped Potentiometers as Accurate Linear Transducers—K. C. Garner. (*Electronic Engrg.*, vol. 33, pp. 32–35; January, 1961.) A method is given for calculating the shunt resistance values required to compensate for the load error, which enables the potentiometer resistance, the minimum input resistance and the load to be independently specified.

621.3.087.4:621.395.625.3:621.318.134 1314
Durable High-Resolution Ferrite Transducer Heads employing Bonding Glass Spacers—S. Duinker. (*Philips Res. Repts.*, vol. 15, pp. 342–367; August, 1960.) The construction is described of various types of ferroxcube magnetic recording heads which have high resistance to wear and improved signal-to-noise ratio.

621.3.087.4:621.395.625.3:681.142 1315
The Development of the Flexible-Disk Magnetic Recorder—R. T. Pearson. (PROC. IRE, vol. 49, pp. 164–174; January, 1961.) The optimum operating conditions for supporting a rotating flexible mylar with disk magnetic coating above a rigid back plate using a gas film are discussed.

621.3.087.4:621.395.625.3:681.142 1316
High-Density Digital Magnetic Recording Techniques—A. S. Hoagland and G. C. Bacon. (PROC. IRE, vol. 49, pp. 258–267; January, 1961.) A comprehensive readback simulation computer program is described which will automatically simulate, for any characteristic pulse, all possible readback signal patterns and test them for specified reading logic as a function of bit density.

621.314.63 1317
Analysis of the D.C. Characteristic of Selenium Rect fiers—H. Lauckner. (*Z. angew. Phys.*, vol. 12, pp. 171–180; April, 1960.) The method of analysis given permits the separation of the various physical phenomena in the rectification process. An empirical equation for a limited section of the rectifier characteristic is derived. The reduction of the slope of the logarithmic I/V characteristic relative to the theoretical value is attributed to the effect of the contact of Se with the n-type intermediate layer of CdSe.

621.316.721.078.3 1318
Constant-Current Supply for Very-High-Resistance Loads—R. W. Haisty. (*Rev. Sci. Instr.*, vol. 31, pp. 1297–1298; December, 1960.) Battery-operated transistorized equipment using a GaAs photoresistor for current control in the range 1–100 μA for any load resistance from 0 to 8×10^8 Ω is described. Measured changes in current with various load conditions do not exceed 1.5 per cent.

681.188:621.395.625.3.001.4 1319
The Testing of Metal Tape Magnetic Cores for Use in a Code Translator—J. D. Andrews. (*Electronic Engrg.*, vol. 33, pp. 21–26; January, 1961.) Acceptance tests which simulate the actual conditions under which the cores will be used are described.

TELEVISION AND PHOTOTELEGRAPHY

621.397:621.391.812.62 1320
Weather Conditions during Television Reception from East Germany and Italy—G. P. A. Braam. (*Tijdschr. ned. Radiogenoot.*, vol. 25, no. 3, pp. 113–129; 1960.) The relation between meteorological conditions and the reception of television signals over exceptionally long distances is investigated. Daily field-strength measurements taken over a period of eight months are discussed with reference to refractive-index gradients computed from radiosonde observations.

621.397:621.391.837.2 1321
The Visual Interference Effect of Horizontal Picture Instability in Television—U. Messerschmid. (*Rundfunktech. Mitt.*, vol. 4, pp. 74–79; April, 1960.) Subjective tests were made to assess the interference caused by horizontal oscillations of the picture during transmissions from video tape recordings or in live transmissions with mains-controlled synchronization. The visual effect is most acute at frequencies between 5 and 15 cps, and the maximum permissible amplitude of horizontal picture displacement is about 0.7 picture elements.

621.397.23:534.76 1322
Stereophonic System for Television Broadcasting—R. B. Dome. (*Electronics*, vol. 33, pp. 71–73; December 16, 1960.) Design considerations are discussed and details given for a system employing a double sideband with suppressed subcarrier to transmit stereo information.

621.397.331.24 1323
Effective Spot Size in Beam Scanning Tubes—A. Sandor. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 69, pp. 735–738; October, 1960.) The influence of scanning velocities, phosphor characteristics and background lighting on the real and apparent reduction in vertical spot size is discussed with reference to measurements made using a cathetometer.

621.397.331.24 1324
Electron Transmission of Mesh Lenses for Scan Magnification in Television Picture Tubes—P. J. Dolon and W. F. Niklas. (*J. Brit. IRE*, vol. 20, pp. 911-919; December, 1960.) A comprehensive mathematical analysis of the electron transmission of mesh lenses is given. Experimental measurement methods are detailed, and the results obtained are compared with theoretical values.

621.397.334.24 1325
Dynamic Spot Formation in Colour Tubes—A. Sandor. (*J. Soc. Mot. Pict. Telev. Engrs.*, vol. 69, pp. 738–742; October, 1960.) An investigation of the spot size requirements in various color-tube systems and a critical assessment of the Lawrence system in its original

form and with different scan-to-grid relations is given.

621.397.61 1326

The Spectral Composition of the Statistical Fluctuations in Present-Day Television Camera Installations—H. Fix and A. Kaufmann. (*Rundfunktech. Mitt.*, vol. 4, pp. 60–65; April, 1960.) The spectral composition of statistical fluctuations is measured on different types of camera tube with or without aperture correction. From this an assessment of the effective signal-to-noise ratio is made. See also 2897 of 1956 (Theile and Fix).

621.397.61 1327

Preamplifiers for Vidicon Cameras using Drift Transistors—H. Anders. (*Rundfunktech. Mitt.*, vol. 4, pp. 66–73; April, 1960.) An experimental amplifier circuit is given and the choice of input circuit and selection of working point for optimum signal-to-noise ratio are discussed. Noise measurements were made on a number of transistors of different type and make. The spectral distribution of statistical fluctuations is compared with that of a tube amplifier and the visual-interference effects are evaluated. Temperature dependence of operating characteristics and means of minimizing it are also considered.

621.397.61:621.396.677 1328

The Use of a High-Gain Television Transmitting Aerial in a Populous Area—Monteath, Millard and Whythe. (See 1074.)

621.397.62 1329

Methods for Perfect Black Reproduction in Television—H. Grosskopf. (*Radio Mentor*, vol. 26, pp. 041–043; January, 1960.) The means of achieving close agreement between the black level of the receiver and that of the transmission monitor are discussed. A simple test signal is suggested as a basis for a subjective assessment and for the adjustment of the black level of the receiver. See also 1377 of 1959.

621.397.74 1330

Recent Methods and Results of Television Network Planning—H. Eden, H. W. Fastert, and K. H. Kaltbeitzer. (*Rundfunktech. Mitt.*, vol. 4, pp. 41–47; April, 1960.) A method of determining idealized networks with minimum mutual interference between transmitters is proposed which includes the process of "density adaptation" to suit local differences in transmitter distribution. The application of the method to the preparation of frequency-allocation plans is discussed. For the English version see *E.B.U. Rev.*, no. 60A, pp. 54–59; April, 1960. See also 717 of February (Correction, *ibid.*, insert).

621.397.74 1331

Transmitter Networks with Nonlinear Channel Distribution—F. Maarleveld. (*Rundfunktech. Mitt.*, vol. 4, pp. 57–59; April, 1960.) The use of nonlinear channel distribution in television network planning is considered; it may provide more favorable results than network planning based on linear distribution (see *e.g.*, 1330 above). For the English version see *E.B.U. Rev.*, no. 60A, pp. 70–72; April, 1960.

TUBES AND THERMIONICS

621.382.002.2 1332

The Diffused Shot-Melting Technique for Making Germanium and Silicon *p-n* Junction Devices—I. A. Lesk. (*J. Electrochem. Soc.*, vol. 107, pp. 534–536; June, 1960.) A description of a simple flexible method for making semiconductor devices is given. Lifetime and resistivity changes which occur during the shot-melting cycle may be minimized by alloy gettering or annealing.

621.382.049.7:537.311.33 1333

Laminar Junction Layers—New Concept in Microcircuits—J. E. Allegretti and D. J. Shombert. (*Electronics*, vol. 33, pp. 55–57; December 2, 1960.) Vapor-phase deposition of single-crystal Si layers is used to form individual semiconductor devices at predetermined locations within the crystal.

621.382.2:621.372.632 1334

Investigations on Semiconductor Mixer Diodes at 3000 Mc/s—G. Salzmann. (*Nachrtech.*, vol. 10, pp. 109–115; March, 1960.) A report on comparative measurements of noise figures and operating characteristics of crystal-diode mixers of East-German and Soviet manufacture, intended for use in waveguide and microstrip circuits, is given.

621.382.23 1335

Shift of Peak Voltage with Temperature in Tunnel Diodes—R. P. Nanavati. (PROC. IRE, vol. 49, p. 349; January, 1961.) Reduction to liquid-nitrogen temperatures caused the peak voltage to increase by 15 to 70 per cent of the value at room temperature.

621.382.23 1336

Capacitance of *p-n* Junctions at Low Temperatures—B. M. Vul and É. I. Zavaritskaya. (*Zh. Eksp. Teor. Fiz.*, vol. 38, pp. 10–17; January, 1960.) Results of capacitance and Q-factor measurements on Ge and Si diodes down to liquid-He temperatures and frequencies up to 1 Mc are detailed. All the phenomena observed can be explained using a simple equivalent circuit.

621.382.23:621.372.44 1337

The Design of Varactor Diodes—J. Hilibrand and C. F. Stocker. (*RCA Rev.*, vol. 21, pp. 457–474; September, 1960.) A figure of merit in terms of sub-harmonic-oscillator circuit parameters is suggested. Optimized parameters are given for the design of diodes for forward- and reverse-bias operation.

621.382.23:621.372.632 1338

Experimental Tunnel-Diode Mixer—J. C. Greene and E. W. Sard. (PROC. IRE, vol. 49, pp. 350–351; January, 1961.) The measured noise factor of a mixer was found to be in fair agreement with that given theoretically by Breitzer (2929 of 1960), but was larger than that predicted by Chang, *et al.* (2644 of 1960).

621.382.233 1339

Uniform Avalanche Effect in Silicon Three-Layer Diodes—A. Goetzberger. (*J. Appl. Phys.*, vol. 31, pp. 2260–2261; December, 1960.) "Interaction of current gain and avalanche multiplication in three-layer diodes is utilized to produce uniform avalanche effect, indicated by uniform light emission over the area of a junction. Proof that the effect is caused by the three-layer action is furnished by removing the emitter layer, which changes the light emission to the usually observed microplasma pattern."

621.382.3 1340

The Tecnetron as a Circuit Element—A. V. J. Martin. (*J. Phys. Radium*, vol. 21, suppl. to no. 7, *Phys. Appl.*, pp. 113A–122A; July, 1960.) Equivalent circuits of the tecnetron are derived and a study is made of the dependence of its characteristics on frequency.

621.382.3 1341

Introduction to the Theory of the Tecnetron—A. V. J. Martin. (*J. Phys. Radium*, vol. 21, suppl. to no. 3, *Phys. Appl.*, pp. 24A–36A; March, 1960.) The basic theory of the device is developed, and certain secondary effects and their bearing on practical design are discussed.

621.382.3:621.317.7 1342

Transistor Measurements—B. N. Harden and R. W. Smith. (*Electronic Tech.*, vol. 38, pp. 58–62; February, 1961.) "Equipment for the direct measurement, at frequencies from 0.5 to 100 Mc/s, of the power gain and noise factor of transistors operating as small-signal common-emitter amplifiers is described. The results of a few typical measurements are included, made under defined conditions of external feedback."

621.382.3:621.385.3 1343

A Proof of the Complete Analogy between Transistor and Valve Triode—H. Tigler. (*Elektronik*, vol. 9, pp. 79–81; March, 1960.) Operating characteristics for different circuit configurations are compared.

621.382.3.001.4 1344

The Execution and Discussion of Comparative Life Measurements on European Low-Frequency Types of Transistor of Various Makes—J. S. Vogel and M. J. O. Strutt. (*Arch. elekt. Übertragung*, vol. 14, pp. 121–131; March, 1960.) Measurements were made on six equivalent types of Ge transistor. The spread between the data is discussed, and particular reference is made to the "creep" of measured values and the "48-hour effect."

621.382.3.012.8 1345

Examination of the Usefulness of Zawels' Practical Equivalent Circuit—O. Müller. (*Elektron. Rundschau*, vol. 14, pp. 90–94; March, 1960.) Formulas are given for the h and y parameters of Zawels' circuit [see *e.g.*, 364 of January (Benz)]. Good agreement is found between calculated and measured parameters up to frequencies beyond α cutoff.

621.382.333 1346

Potential Distribution and Capacitance of a Graded *p-n* Junction—S. P. Morgan and F. M. Smits. (*Bell Sys. Tech. J.*, vol. 39, pp. 1573–1602; November, 1960.) Formulas, graphs, and tables are given which simplify the solution of the equation for the potential distribution. The expression for the capacitance is split into two parts, one of which dominates in the neutral case and the other in the space-charge case.

621.382.333 1347

Observation of an Anomaly in Transistor Characteristics—O. Nakahara. (*J. Phys. Soc. Japan*, vol. 15, pp. 1537–1538; August, 1960.) An interpretation is given of the origin of the negative collector conductance observed in an alloy *p-ν-p* transistor with base of 40-Ωcm *n*-type Ge. See 3538 of 1958 (Shields).

621.382.333:621.318.57 1348

Switching Times for Alloy Junction Transistors—P. James and A. F. Newell. (*Mullard Tech. Commun.*, vol. 5, pp. 159–171; June, 1960.) The equivalent circuit for large-signal operation of a transistor is derived and used to calculate the transient response of transistors in some typical circuits. Methods of measuring the equivalent-circuit parameters are described.

621.382.333:621.387.4 1349

On the Use of 2N504 Transistors in the Avalanche Mode for Nuclear Instrumentation—R. Fullwood. (*Rev. Sci. Instr.*, vol. 31, pp. 1186–1189; November, 1960.) Circuits are described using the transistors as fast discriminators on photomultiplier tubes in nuclear counters, and as pulse generating elements in rectangular-pulse and double-pulse generators.

621.382.333.4 1350

The Binistor—a New Semiconductor Device—N. DeWolf. (*Electronic Ind.*, vol. 19, pp. 84–87; August, 1960.) A general description of a silicon four-layer device suitable for use in switching and storage circuits is given.

621.383.41 1351
Preparation and Properties of PbTe Photoresistors with Controlled Stoichiometric Composition—K. Gürs. (*Z. Phys.*, vol. 158, pp. 533–552; April 12, 1960.)

621.383.41 1352
Influence of Crystal Size on the Spectral Response Limit of Evaporated PbTe and PbSe Photoconductive Cells—W. D. Lawson, F. A. Smith, and A. S. Young. (*J. Electrochem. Soc.*, vol. 107, pp. 206–210; March, 1960.) By controlling crystallite size, the response or the absorption of a layer can be made to terminate at any wavelength up to the limit fixed by the energy gap of the compound.

621.383.5 1353
Silicon Photovoltaic Cells for Instrumentation and Control Applications—V. Magee and A. A. Shepherd. (*J. Brit. IRE*, vol. 20, pp. 803–819; November, 1960. Discussion.) The operation of *p-n* junction photocells for detection of visible and near-infrared radiation is described. An account of the detection mechanism is given and construction techniques are outlined. Applications described include relay systems, data reading, machine-tool control and binary counting.

621.383.53 1354
Thermal and Optical Behaviour of Phototransistors—D. Shaw and B. Crump. (*Electronic Engrg.*, vol. 32, pp. 753–757; December, 1960.) Investigation of thermal effects on Ge and Si phototransistors and their application to a precise position-measuring device is discussed. Suitable circuits are given for operation of the Ge device at temperatures to 55°C, and Si to about 100°C.

621.385.032.212 1355
Electron Emission from Cold Magnesium Oxide—H. N. Daglish. (*Proc. IEE*, vol. 108, pt. B, pp. 103–110; January, 1961.) Methods of making and processing cathodes of this type are described, and it is shown that stable emission is obtained only from cathodes in a fully oxidized state. The behavior of these cathodes under various experimental conditions has been examined with particular reference to voltage and temperature effects.

621.385.032.213 1356
Space-Charge Neutralization and Thermionic Emission—R. N. Franklin. (*J. Electronics and Control*, vol. 9, pp. 385–390; November, 1960.) An exact solution for the density distribution of space charge and the field is derived. The general shape of the potential distribution can be found as a function of the ratio of positive ions to electrons, and it is shown that the transition from pure electron emission to ion emission is smooth and that no oscillatory solution for the potential exists.

621.385.032.213 1357
Investigation of Electrostatic Emission of Electrons from a Tungsten Emitter in Pulse Operation—I. I. Gofman, O. D. Protopopov, and G. N. Shuppe. (*Fiz. Tverdogo Tela*, vol. 3, pp. 1323–1327; June, 1960.) Measurements have been made of I/V characteristics and the results are compared qualitatively with theory.

621.385.032.213.13 1358
Temperature Variations of an Oxide Cathode produced by the Flow of Thermionic Current—G. Mesnard and R. Uzan. (*Le Vide*, vol. 15, pp. 301–312; July/August, 1960. In French and English.) Results show that temperature variations are different for the core and the external coating of the cathode and that further temperature changes occur during the life of the tube. The mechanism of electron transfer through the coating is found to be strongly influenced by porosity and residual gases.

621.385.032.213.13 1359
Emission from Miniature Hollow Cathodes—A. Sandor. (*Proc. IEE*, vol. 108, pt. B, pp. 90–96; January, 1961.) A detailed study of the properties of cathodes internally coated with conventional oxides, with special reference to the need for higher current densities and a smaller optical source, is given.

621.385.032.213.13 1360
Pressed-Oxide Nickel-Matrix Cathode below Apertured Electrodes—A. Sandor. (*Proc. IEE*, vol. 108, pt. B, pp. 97–102; January, 1961.) The advantages and construction of this type of high-temperature cathode are outlined, and performance is discussed in detail. Excessive vapor generation limits its usefulness when narrow electrode apertures are used.

621.385.032.213.6 1361
Some Electrical and Surface Properties of Sintered Nickel Matrix Cathodes—R. W. Fane. (*Brit. J. Appl. Phys.*, vol. 11, pp. 513–516; November, 1960.) Measurements of the work function of Ni cathodes indicate values in the region of 1 ev. Schottky plots are found to agree with theory, provided the cathode surface is polished. For preparation and metallurgy of Ni cathodes see 2600 of 1958 and 4250 of 1959 (Boone, *et al.*).

621.385.032.269.1 1362
Determination of Electrode Shapes for Axially Symmetric Electron Guns—K. J. Harker. (*J. Appl. Phys.*, vol. 31, pp. 2165–2170; December, 1960.) An analytic continuation of Laplace's equation into a fictitious complex domain makes it suitable for application to a space-charge-limited curvilinear-flow electron gun. The method should be useful in applications where the alternative solution is to use analogs such as the electrolyte tank.

621.385.1:621.375.2 1363
Special Amplifier Valves and their Characteristics: Survey of Types and Examples of Circuits—W. Geist. (*Elektron. Rundschau*, vol. 14, pp. 129–132. 136; April, 1960.) Data tables covering tubes designed for special applications are given.

621.385.6 1364
Kinematic Electron Bunching by Sinusoidal Travelling and Standing Waves in Short Extended Interaction Regions—H. Golde. (*J. Electronics and Control*, vol. 9, pp. 285–302; October, 1960.) The treatment follows the general theory developed by Wessel-Berg and is applied to the large-signal condition. The results agree well with those obtained by numerical integration using a computer.

621.385.6:537.533 1365
In What Sense do Slow Waves Carry Negative Energy?—P. A. Sturrock. (*J. Appl. Phys.*, vol. 31, pp. 2052–2056; November, 1960.) The positive and negative energy, respectively, of fast and slow space-charge waves, as found in electron tubes and proved by the "small-amplitude power theorem," is derived in a general way starting with a quadratic Lagrangian function.

621.385.6:537.533 1366
Conservation Laws of an Interacting Electron Beam—M. C. Pease. (*J. Appl. Phys.*, vol. 31, pp. 2028–2036; November, 1960.) The linearized theory of electron beams interacting with a circuit is studied. Methods are obtained for determining, from the differential equations of the system, the quadratic invariance, if it exists, which is independent of a given parameter. This leads to a modal description and a kinetic power law, such as that of Chu.

621.385.6:537.533 1367
Oscillations in Long Electron Beams—B. Agdur. (*Ericsson Tech.*, vol. 16, no. 1, pp. 43–57; 1960.) Two types of instability in long electron beams confined by magnetic fields are investigated. One, which is dependent on the presence of positive ions trapped in the beam, occurs at frequencies near the plasma frequency of the ions and gives rise to a rotational motion of the electron beam. The other, which is dependent on the presence of slow electrons in the beam, occurs at frequencies close to the gyro-frequency of the electrons.

621.385.6:537.533:621.391.822 1368
Noise Propagation in Drifting Multivelocity Electron Beams—J. A. Morrison. (*J. Appl. Phys.*, vol. 31, pp. 2066–2067; November, 1960.) Some of the macroscopic formulas of Berghammer and Bloom (2565 of 1960) do not agree with the microscopic numerical calculations based on the distribution-function method of Siegman, *et al.* (749 of 1958).

621.385.6:621.375.9:621.372.44 1369
Transverse Electron-Beam Waves in Varying Magnetic Fields—E. I. Gordon. (*Bell. Sys. Tech. J.*, vol. 39, pp. 1603–1616; November, 1960.) The cyclotron wave associated with the rotational motion of the beam electrons and the synchronous wave associated with the spatial configuration of the beam are shown to be coupled by varying magnetic fields.

621.385.6:621.375.9:621.372.44 1370
An Electron-Beam Parametric Amplifier for the 200-Mc/s Region—G. O. Chalk. (*Proc. IEE*, vol. 108, pt. B, pp. 125–132; January, 1961.) The operation of the Adler tube [see, *e.g.*, 361 of 1960 (Adler, *et al.*)] is outlined. The construction of a tube giving a noise figure of 1.6 db, 10 db gain and bandwidth 25 Mc is given.

621.385.63 1371
Experimental Travelling-Wave Tubes—G. M. Clarke. (*Electronic Tech.*, vol. 38, pp. 38–45; February, 1961.) The design and performance of tubes with CW outputs up to 300 w and gains greater than 20 db are described.

621.385.63 1372
The Magnetic Field and Flux Distributions in a Periodic Focusing Stack for Travelling-Wave Tubes—M. J. Schindler. (*RCA Rev.*, vol. 21, pp. 414–436; September, 1960.) The problem is treated as a static one, rather than by the usual electrodynamic analogy, taking account of the charges on the disk and outer hub surfaces, and the effect of neighboring magnets. Measurements verify the validity of the approach, and design formulas are given.

621.385.632.14 1373
A Corrugated Bent Waveguide as a Delay System with Positive Dispersion—Z. I. Taranenko. (*Izv. vyssh. uch. Zav., Radiotekhnika*, vol. 3, pp. 30–39; January/February, 1960.) An examination is made of the dispersion curves for an electron beam in a slow-wave structure comprising a rectangular waveguide: a) with equal transit apertures, and b) with apertures of two different diameters arranged alternately. Formulas are derived for the effective impedance of the system.

621.385.633 1374
Operation of the Backward-Wave Oscillator—M. Y. Wong, G. D. Sims and I. M. Stephenson. (*Nature*, vol. 188, pp. 803–804; December 3, 1960.) The traveling-wave-tube analog described earlier (1078 of 1960 (Sims and Stephenson)] has been operated as a backward-wave oscillator and measurements have been made of the voltage distribution along the slow-wave circuit. These indicate that at least

a 3-mode theory is necessary to explain the operation of backward-wave oscillators.

621.385.64 1375

General Steady-State Theory of Cylindrical Magnetrons—P. A. Lindsay. (*J. Electronics and Control*, vol. 9, pp. 241–283; October, 1960.) The theory takes into account space charge and the Maxwellian distribution of emitted electrons, but collision effects are neglected. Calculations are given of the volume density of the electrons and the radial and azimuthal components of the current density at any point between cathode and anode.

621.385.644 1376

An Experimental Wide-Tuning-Range Inverted Magnetron—A. Singh and N. C. Vaidya. (PROC. IRE, vol. 48, pp. 2035–2036; December, 1960.) Constructional details and performance data are given for an inverted interdigital magnetron tuned by a coaxial line. See 4066 of 1957 (Singh).

621.385.644 1377

The Resonator and Output Couplers for a Wide-Tuning-Range Inverted Magnetron—A. Singh, N. C. Vaidya, R. A. Rao, K. Chandra, and G. S. Siddhu. (*J. Inst. Telecommun. Engrs., India*, vol. 6, pp. 153–160; June, 1960.) Measurements were made on an inverted interdigital magnetron. With an iris coupling the resonator to the output line, uniform external Q was obtained over a 2:1 band. See also 1381 above.

621.385.832.032.265:621.317.755 1378

Velocity Modulation and Related High-Frequency Deflection Errors of the Travelling-Wave Deflection System—J. Goldberg. (*Rev. Sci. Instr.*, vol. 31, pp. 1320–1325; December, 1960.) A mathematical analysis is given of the potential and electric-field distributions. Generalized expressions are derived for the errors in beam deflection due to nonuniformity and dispersion.

621.385.832.032.269:621.317.329 1379

Method for Theoretical Investigation of Type-O Focusing—J. E. Picquendar and O. Cahen. (*Rev. tech. Comp. franç. Thomson-Houston*, no. 32, pp. 7–39; February, 1960.) Curves, tables and simplified equations are given for calculating electron-gun perveance and tracing electron-beam trajectories under the influence of an axial magnetic field.

621.387:621.362 1380

Potential Distributions in a Low-Pressure Thermionic Converter—Auer. (See 1279.)

621.387:621.362 1381

Experimental Investigations of the Cesium Plasma Cell—Ranken, Grover, and Salmi. (See 1280.)

621.387.132.223:621.396.96 1382

Gas Clipper Tubes for Radar Service—W. W. Watrous and J. McArtney. (*Electronics*, vol. 33, pp. 80–83; December 16, 1960.) Hydrogen-filled thyratrons prevent the building up of dangerous voltages in the pulse-forming circuits.

Membership

(Continued from page 92A)

Ormston, W. J., Nutley, N. J.
Otero, J. L., Tucson, Ariz.
Oxendale, J. L., Victoria, B. C., Canada
Parr, J. C., Elmhurst, L. I., N. Y.
Pezirtzoglou, E. I., Mountain View, Calif.
Philipson, R. B., Merchantville, N. J.
Pisarski, J., New York, N. Y.
Pongracz-Bortha, E., Van Nuys, Calif.
Prescott, H., Washington, D. C.
Prickett, C. E., Pomona, Calif.
Provencher, J. H., San Diego, Calif.
Porter, W. A., Ann Arbor, Mich.
Radsliff, J. L., Topeka, Kan.
Ramos, D., Bryan, Tex.
Raney, A. B., Fort Worth, Tex.
Raymond, H. G., Bedford, Mass.
Reilly, F. N., Richardson, Tex.
Roberts, R. W., Jr., Sunnyvale, Calif.
Robins, K. E., Dayton, Ohio
Robinson, L. A., Mountain View, Calif.
Rode, J. J., Leawood, Kan.
Rodgers, J. C., Baltimore, Md.
Romagnoli, J. D., Haddonfield, N. J.
Romanchick, M., Chicago, Ill.
Roscoe, B. J., Baltimore, Md.
Rosenthal, H., Levittown, Pa.
Ruttenberg, S. A., Los Angeles, Calif.
Ryan, E. D., Downey, Calif.
Sagnard, F. R., Paris, Seine, France
Scent, R. H., St. Charles, Ill.
Schlueter, R. C., Collingswood, N. J.
Schmidt, P. S., Houston, Tex.
Schmoock, R. F., Ivyland, Pa.
Schofield, A., Leeds, England
Sha, R. T., Boonton, N. J.
Shaw, R. S., Passaic, N. J.
Sheng, A. S., Haddonfield, N. J.
Sherman, B. S., New York, N. Y.
Silberbach, R. O., Chicago, Ill.
Singleton, J. W., Jr., Baltimore, Md.
Skaggs, J. A., Santa Monica, Calif.
Slabaugh, J. E., Dallas, Tex.
Smith, P. L., Jr., Pittsburgh, Pa.
Sparks, C., Jr., Van Nuys, Calif.
Spear, A. F., Birmingham, Ala.
Spraker, M. R., Palo Alto, Calif.
Squire, F. C., Jr., Mount Holly, N. J.
Stevens, E. J., Clemson, S.C.
Stiebel, R., Chicago, Ill.
Stokes, J. D., Jr., Washington, D. C.
Stone, W. A., Richardson, Tex.
Templin, J., Prairie Village, Kan.
Thomas, C. M., Pomona, Calif.
Thompson, G. W., U. S. Air Force Academy, Colo.
Tingleff, R. D., Granada Hills, Calif.
Tiuri, M. E., Helsinki, Finland
Toft, P. D., Oxnard, Calif.
Tomlinson, R. B., Deep River, Ont., Canada
Toomay, J. C., Griffiss AFB, N. Y.
Torf, A. S., Fort Ritchie, Md.
Torrero, E. A., Brooklyn, N. Y.
Tossell, A. R., Baltimore, Md.
Tuckerman, R. G., Damascus, Md.
Tubman, V. H., Victoria, B. C., Canada
Unbehauen, R., Stuttgart-Schonberg, Germany
Van Der Poel, J. M., East Paterson, N. J.
Van Peer, N. J., Chicago, Ill.
Ward, P. W., Dallas, Tex.
Wartenberg, M. R., Brooklyn, N. Y.
Weiss, C. M., Riverton, N. J.
Wetherbee, J. K., Columbus, Ohio
Whittier, F. W., Swarthmore, Pa.
Wieman, F., Seattle, Wash.
Williams, G. L., Torrance Calif.
Williamson, R. H., Tujunga, Calif.
Wilson, C. E., King of Prussia, Pa.
Wilson, Q. C., Los Angeles, Calif.

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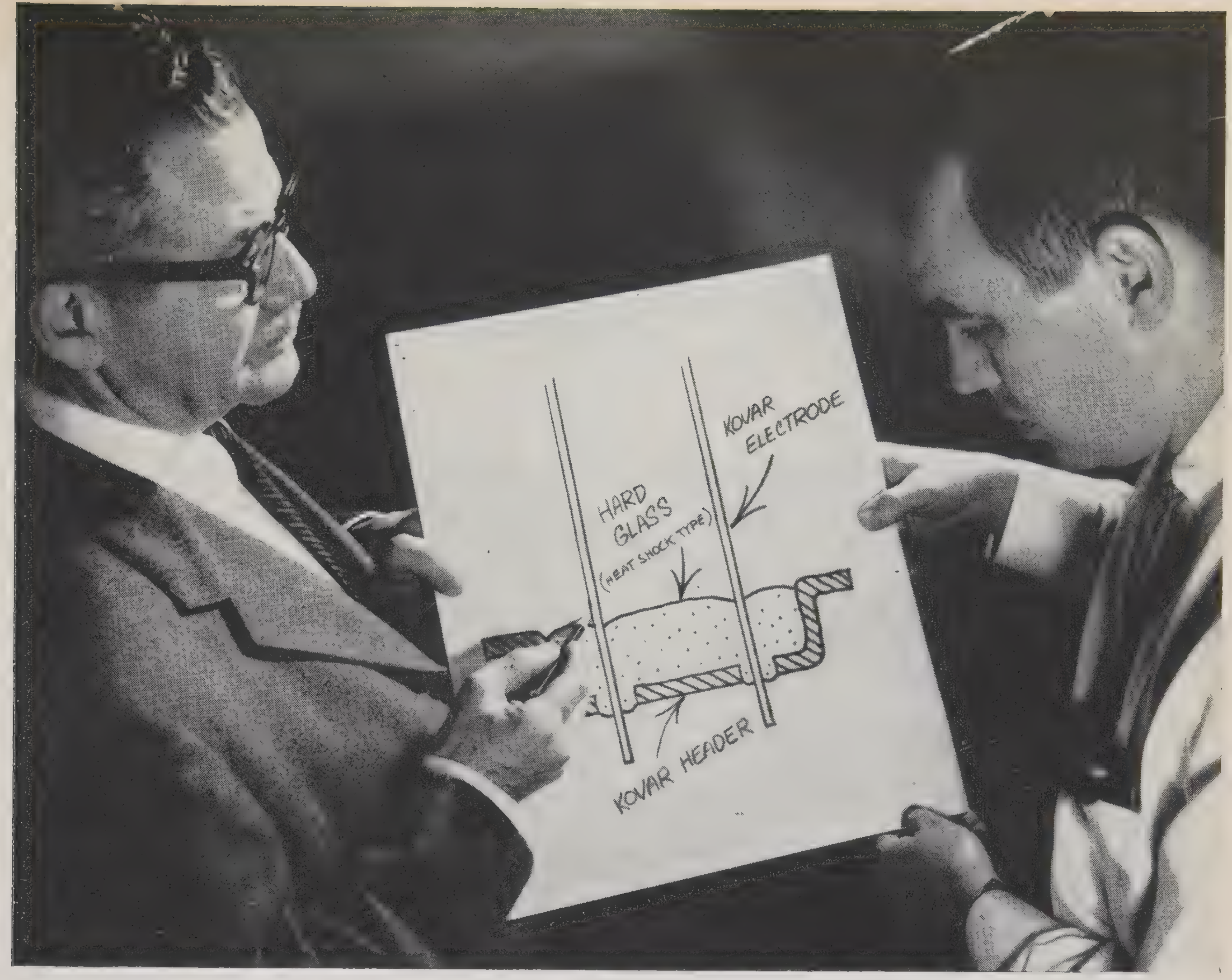

KOVAR
ELECTRODE
HARD
GLASS
(HEAT SHOCK TYPE)
KOVAR HEADER

Membership

(Continued from page 98A)

Wilson, W. D., Overton, Neb.
Wong, D. J., Poughkeepsie, N. Y.
Woodhull, J. H., Silver Spring, Md.
Wright, B. K., Palo Alto, Calif.
Yopp, W. B., Inglewood, Calif.
Zampier, A. L., New Hartford, N. Y.
Zelesky, F., Norwalk, Conn.
Ziegler, G. C., Greenwich, Conn.

Admission to Associate

Alberts, A. R., Newton Highlands, Mass.
Angus, J. G., Schenectady, N. Y.
Annett, M. C., Berwyn, Ill.
Baker, T. L., Los Angeles, Calif.
Becker, S. M., Red Bank, N. J.
Becker, V. H., Colfax, Calif.
Bernstein, R. K., New York, N. Y.
Brooks, G. W., Owensboro, Ky.
Buchanan, E. J., Sanford, N. C.
Busby, J. M., Huntsville, Ala.
Capocaccia, F. E., Milan, Italy
Caspi, E., Tel Aviv, Israel
Castelli, A. L., Milan, Italy
Daniels, I. D., Owensboro, Ky.
Davis, D. K., La Puente, Calif.
Dean, B. J., Ottawa, Ont., Canada
Dolazza, E., Milan, Italy
Donovan, K. H., APO, New York, N. Y.
Ellis, R. N., Glens Falls, N. Y.
Epstein, S., Paramus, N. J.
Essington, L. B., San Antonio, Tex.
Feibel, S. J., Dorchester, Mass.
Fenske, E. R., Alexandria, Va.
Fitch, E. S., Baltimore, Md.
Fulchiero, E., New London, Conn.
Funkhouser, J. R., Burlington, Iowa
George, C., Manhasset, L. I., N. Y.
Geraci, P. C., Hyattsville, Md.
Goodrich, F. M., Redwood City, Calif.
Grimes, D. W., La Vale, Md.
Henderson, J. G., Baltimore, Md.
Heroux, N. M., Watertown, Conn.
Hilderbrand, H. H., Geneva, Switzerland
Holt, R. A., Lexington, Mass.
Iseri, V. C., Los Angeles, Calif.
Ives, R. A., Tonawanda, N. Y.
Janyska, W. F., Baltimore, Md.
Judevine, D. E., Philadelphia, Pa.
Kent, L. G., Utica, N. Y.
Kircher, W. L., Glen Oaks, N. Y.
Larson, R. P., Cedar Rapids, Iowa
Laughlin, J. L., Hinsdale, Ill.
Lauterer, P. W., Toronto, Ont., Canada
Lynn, F. H., Whitesboro, N. Y.
Mahoney, D. E., Inglewood, Calif.
Mather, J. B., Alhambra, Calif.
McBride, H. E., Anchorage, Alaska
McNamara, J. D., Roslyn, Pa.
Meyer, F. S., Utica, N. Y.
Miles, H. H., Washington, D. C.
Mitchell, C. H., Little Rock, Ark.
Myers, H. M., Lynn, Mass.
Novak, J. L., Manhattan Beach, Calif.
O'Neill, D. R., Haddonfield, N. J.
Payne, G. R., Midwest City, Okla.
Penning, D. C., Menlo Park, Calif.
Phinney, S. E., Boston, Mass.
Quagliana, V. E., Buffalo, N. Y.
Quane, E., Dearborn, Mich.
Quenot, J., Brunoy, France
Reed, K. A., Poughkeepsie, N. Y.
Reitz, W. L., Denver, Colo.
Richardson, R. G., Minneapolis, Minn.
Riegel, R. E., Orlando, Fla.
Rollf, R. K., Bedford, Mass.
Ruhl, G. L., Hawthorne, Calif.
Sanders, I. W., Jr., Sylmar, Calif.
Sigal, B. M., North Hollywood, Calif.
Spencer, S. B., Victoria, B. C., Canada
Strelczyk, E. S., Waterford, Conn.

(Continued on page 102A)

Membership

(Continued from page 100A)

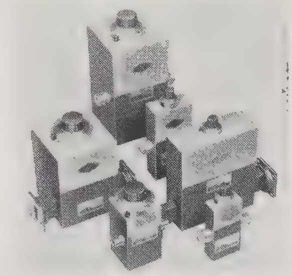
Tan, P. K., Ardmore, Auckland, New Zealand
Vollmer, C. B., Sr., Monclova, Ohio
Wall, A., Waltham, Mass.
Wayne, B. C., Encino, Calif.
Wells, R. H., Baltimore, Md.
Willard, D. U., Baltimore, Md.
Williams, J. T., St. Louis, Mo.
Wise, M., La Puente, Calif.
Wong, H. W., Syosset, L. I., N. Y.
Woodruff, W. W., Great Lakes, Ill.
Young, R. M., Waltham, Mass.
Zeidlhack, D. F., Portland, Ore.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

Microwave Tunable Band-Pass Filters

The John Gombos Co., Inc., Webro Rd., Clifton, N. J., has recently announced the availability of their complete line of microwave tunable band-pass filters from 500 to 18,000 mc. The units feature 2-, 3-, and 4-section cavities, offering a selection of rejection characteristics, and low insertion loss.

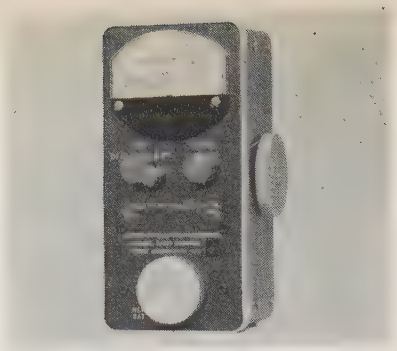


There are 81 models to select from according to specific requirements, covering UHF, L, S, C, X, and Ku Bands. These models are designed for laboratory applications such as image filters, noise measurements, harmonic rejection, frequency selection, and spectrum analysis. They are also designed for systems applications such as video or superheterodyne preselection, resonant diplexing, frequency discrimination, multiplexing and band elimination.

Data sheets for each tunable filter are available, along with detailed information on special applications, by writing to the firm.

Vibration Calibrator

A compact battery-operated vibration calibrator (Type 1557-A), for calibrating accelerometers, vibration pickups and vibration meters, as well as other vibration-measurement systems that use small, crystal-type accelerometers as sensing elements, has been announced by the **General Radio Co.,** West Concord, Mass.



Consisting of a transistorized electromechanical oscillator and a battery-operated cylindrical shaker, the instrument provides a standard acceleration for 1 g (*rms*) at 100 cps. Acceleration output appears at two pill-box-shaped 50 gram discs mounted on an internal cylinder extending through the instrument and projecting at the sides. The pickup to be calibrated is attached to, or in place of, one of these discs. Acceleration accuracy is ±10 per cent; frequency accuracy is ±1 per cent.

Oscillator amplitude can be adjusted by a panel level control, and in this way, the calibrator is adapted to pickups of different masses. To subject a pickup to an acceleration of 1 g at 100 cps, the operator adjusts the level control until the panel meter, calibrated in grams, indicates the mass of the pickup. The only other control on the instrument is a combination on-off switch and battery checker.

Power is supplied by four RM-Y mercury cells, whose life is 100 hours.

The total weight of the instrument, including a leather carrying case, is $3\frac{3}{4}$ pounds. Dimensions are 4″×8″×4″.

Price of the 1557-A is $225.00 net, f.o.b., West Concord, Mass.

Packaged Oscillator

A new 1000 kc packaged crystal oscillator with transistorized circuitry is now in production by **Bliley Electric Co.,** Union Station Bldg., Erie, Pa.

The unit is designed for use in frequency counters or as a master oscillator in frequency control systems. This plug-in package is supplied with a high precision glass crystal at 1000 kc and has frequency stability of one part in one hundred million under room ambient temperature conditions. The oscillator can be custom-designed for other frequencies in a range from 950 to 3000 kc. Prototype price approximately $250.00.

(Continued on page 104A)

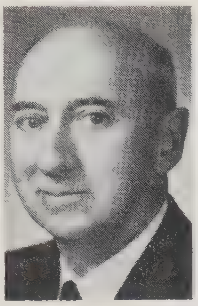
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(*Continued from page 102A*)

J. P. Smith, Jr., Joins Ortho Filter

George G. Pagonis, president of **Ortho Filter Corp.,** a division of **Ortho Industries, Inc.,** 7 Paterson St., Paterson, N. J., manufacturers of electric wave filters, toroidal transformers, magnetic amplifiers, delay lines, equalizers and attenuators, announces that J. P. Smith, Jr., has joined the company as a senior engineer.

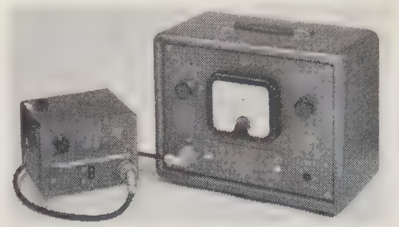


Smith was previously associated with The Daven Co. for over 25 years as chief engineer and director of engineering. A graduate of the University of Arkansas, Smith was formerly affiliated with Lear Radio, Inc., Hygrade Sylvania, and the General Electric Co.

A resident of West Allentown, Pa., Smith is a fellow of the Audio Engineering Society.

Electrometer

The Herman H. Sticht Co., Inc., 27 Park Pl., New York 7, N. Y., has available another instrument in the line of "VIBRON" Electrometers as manufactured by **Electronic Instruments, Ltd.,** Richmond, Surrey, England.



This new instrument is the Model 33C "VIBRON" Industrial Electrometer with Model B33C Converter Unit. It is described in Bulletin #1038. This instrument is a special type of vibrating capacitor electrometer suitable for industrial applications and routine laboratory measurements. It consists of two separate units, the Model 33C which is a high gain amplifier and the Converter Unit B33C which contains the vibrating capacitor unit together with pre-amplifier and input resistors.

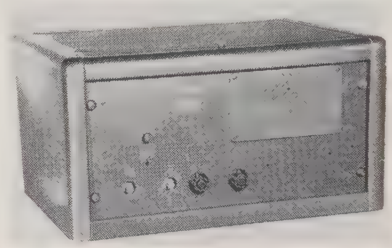
Input ranges of the 33C are 10, 30, 100, 300 and 1000 millivolts. Current measurements as low as 10^{-15} amperes are possible. Resistance measurements as high as 10^{16} ohms can be made. A feature of the instrument is its zero stability which is ± 100 microvolts for a 12 hour period or ± 300 microvolts for one week when fully stabilized. Accuracy of the instrument as a millivoltmeter is $\pm 1\%$ on all ranges.

Bearing-Distance-Heading-Indicator

Newly printed catalog page contains illustration, outline drawing, schematic and complete technical details including specifications, component parts and applications of 3-in-1 BDH Indicator which provides on a single indicator face, information on 2 relative bearings, distance and magnetic heading. Catalog page available without charge from **John Oster Manufacturing Co., Avionic Div.,** Racine, Wis.

Linear Amplifier

Hamner Electronics Co., P.O. Box 531, Princeton, N. J., manufacturer of nuclear instruments, has announced a new double delay line non-overloading linear amplifier, the N-308, based on the Oak Ridge National Laboratories A-8 design. The N-308 is also available with integral or differential discriminator as well as an optional pick-off circuit for coincidence work in the 50 mμs range with sodium iodide.

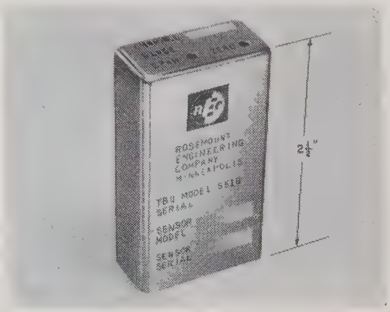


The A-8 amplifier design is suitable where very large overload signals may be present, where counting rates up to 250 kc may be encountered or where it is required to do medium fast coincidence work on pulses having slow rise times.

The Hamner N-308 has a gain of 7000, adjustable by coarse and fine gain controls, stability of 0.25% per day, rise time about 0.2 μs and integral linearity of 0.1%, 3–100 volts into a 10 k or greater load.

Resistance Bridge

A resistance bridge which, for normal applications, is able to suppress the effects of variable and unknown lead resistance better than previously existing methods has been announced by **Rosemount Engineering Co.,** 4900 W. 78th St., Minneapolis 24, Minn.



(*Continued on page 106A*)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 104A)

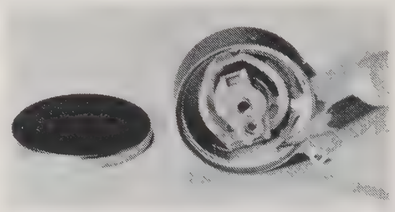
The newly developed variation of the basic Wheatstone bridge was designed for use in temperature probes. With conventional 3-wire and 4-wire bridge circuits, variable and unknown lead resistances can cause prominent errors, particularly in unbalanced conditions. This Triple-bridge suppresses lead resistance effectively both at null and when unbalanced. Typically, when leads vary from 0 to 5 ohms collectively or individually, the effect will not exceed 0.1% of full scale.

The Triple-bridge is a plug-in unit permitting convenient change of full scale temperature and capable of correcting known calibration errors of the temperature probe with which it is used. A basic 10-channel Triple-bridge unit is offered, which contains sockets and inter-connecting wiring for 10 probes and 10 plug-in Triple-bridge units, thus providing 10 temperature ranges for each probe. Optional auxiliary equipment includes a precision regulated power supply and calibration devices for bridge units, power supply and indicators.

For additional information write for Bulletin 86012, to the firm.

12½ Watt Resistor

Military applications for a wire-wound control have led to an increase of more than triple the original wattage rating within the same physical size. The use of special insulating materials by the **Mallory Controls Co.,** Frankfort, Ind., made this upgrading possible in its new "MG" control, which is rated at 12½ watts at an ambient temperature of 40°C.



More economical than a rheostat of the same wattage, the new wire-wound control is derated to zero watts at 300°C, as compared to 340°C for similar derating of a 12½ watt rheostat. The "MG" design is being used in a multi-section control for missile launching ground support equipment and in a temperature control for aircraft cabins.

The "MG" control was developed by Mallory Controls, a division of P. R. Mallory & Co., Inc.

Switch Catalog

The Daven Company, Livingston, N. J., a subsidiary of General Mills, Inc., has just announced the publication of a

(Continued on page 108A)

PORTABLE KLYSTRON POWER SUPPLY 809-A

featuring: • New compact size: 8″ x 12″ x 15″ • New low in reflector voltage ripple: less than 1 mv rms • New planetary gears to give finer adjustment of reflector voltage • New design including internal blower, built-in cabinet tilt stand, PRD expansion coil cord with polarized ac plug • Direct reading of beam voltage or current on front panel meter.

Regulated beam voltage 250 to 600 volts; regulated reflector voltage 0 to −900 volts; 6.3 volt ac filament supply. Reflector voltage available either unmodulated or internally modulated by square wave or sawtooth. Send for data! **PRD ELECTRONICS, INC.:** 202 Tillary St., Brooklyn 1, New York, ULster 2-6800; 1608 Centinela Ave., Inglewood, California, ORegon 8-9048. *A Subsidiary of Harris-Intertype Corporation.*

HARRIS INTERTYPE CORPORATION

New from PRD!

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 106A)

new 48-page catalog of its switch line.

The catalog is available, without charge, to anyone involved in the selection and ordering of switches.

Described and illustrated in the catalog are standard switches such as single deck, shorting; single deck, non-shorting; multiple deck, shorting; multiple deck, non-shorting; as well as special switches like miniature ceramic switches, sub-miniature switches, pre-wired switch assemblies, terminal board switches, solenoid-operated switches, adjustable stop switches, progressive shorting-type switches, high speed switches, commutator-type switches, hermetically-sealed switches and spring return switches.

A complete list of all Daven regional sales engineers and distributors concludes the catalog.

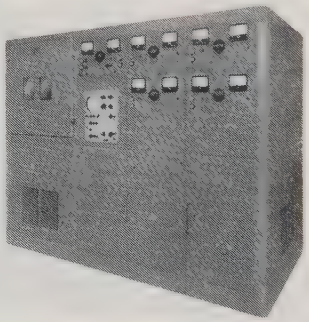
For a copy write to Mr. E. L. Grayson at the firm.

Power Supply Data Sheets

A new series of high power electronic data sheets is now being issued by **FXR, Inc.**, 25–26 50th St., Woodside 77, N. Y., through the company's **High Power Electronics Div.** Under the category of high voltage regulated power supplies, three sheets illustrate and give full technical specifications for Model Z850A, 0–12 kv main supply with optional focus and filament supplies; Model Z852A, 0–18 kv main supply with optional focus, magnet, and filament supplies. The fourth sheet covers the 50 megawatt *S*-band klystron transmitter as well as its remote control console. All sheets are coded with reverse block headings, by instrumentation types, for ease of filing and category identification. These data sheets are available on request by specifying the model number.

Power Triodes and Tetrodes Test Set

The **High Power Electronics Div., FXR, Inc.**, 25-26 50th St., Woodside 77, N. Y., has just released details on their new, 5 Megawatt Positive Grid Region Test Set. This test set can be used for the design, development and production quality control testing of high power triodes and tetrodes whose test operating parameters fall within the wide ranges offered. All voltages necessary for the testing of these tubes are produced internally in the test set. Full protection of both the operating personnel and the components undergoing tests is assured by use of interlocks, overload devices and lead shielding. As a further safeguard, a "panic button" is strategically located.



The test set has built-in provisions for both water and air cooling. Peak grid pulse of the set is 10 kv, 500 amperes at 2 μs. Pulse repetition rate is adjustable from "one shot" to 600 pps. Built-in power supplies include the grid supply detailed above, the plate supply: 30 kv at 200 ma; and the filament supply: 15 volts at 333 amperes, or 30 volts at 167 amperes. All supplies are controllable from the front panel, are adjustable to zero and for test tube protection are sequence locked. A 75 kw forced-air-cooled, non-inductive resistor is provided as a plate load for the tube under test.

Literature is available. For further information as to custom requirements, contact the Sales Engineering Dept.

Spring-Energized Gyro

A new low cost spring-energized gyro has been designed to provide an inertial reference for short duration missiles and other ballistic devices.

(Continued on page 110A)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

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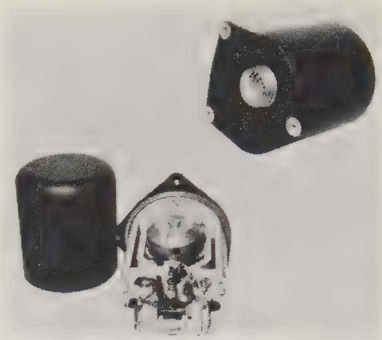


Called Model 1091 by the **Courter Products Div., Model Engineering and Manufacturing Corp.,** (MEMCOR), Boyne City, Mich., the entire self-contained package weighs about $3\frac{3}{4}$ pounds and measures 5 inches long by 4.2 inches in diameter.

The gyro has segmented pickoff commutators on its outer gimbal for roll reference information.

Unlatching is via either one or two 28vdc dimple motors, followed automatically by uncaging which takes a maximum of 100 ms.

Full speed of 3330 RPM is attained in 100 ms, and run-down time is 7 to 9 minutes. The 1091 gyro achieves an angular momentum of 737,000 gm cm²/s with a moment of inertia at 2116 gm cm².

Drift measured after 30 seconds of operation is less than 1° for the outer gimbal and less than 6° for the inner gimbal. Both gimbals have 360° freedom.

The hermetically sealed gyro is said to give full performance for a minimum of 100 firings, and may be used in any application requiring a short duration mechanically energized gyroscopic component. Variations in winding methods may be incorporated when the gyro is to be used repeatedly.

Miniature Plug Catalog

A new catalog has recently been made available by **Cannon Electric Co.,** 3208 Humboldt St., Los Angeles 31, Calif. The catalog presents a thorough and comprehensive description of the KQ/KR line of general purpose miniature plugs designed especially to meet the more severe requirements of aircraft and missile applications. The KQ/KR series features a universal receptacle which will accept either the ball-lock plug or the bayonet-lock plug. All plugs in the series have a moisture sealing feature that eliminates the need for potting, crimp-type snap-in contacts that eliminate the need for soldering, probe-proof socket contacts, polarization by multiple keyways that provide easy mating in blind locations, and the high altitude requirement of MIL-C-26500. Hermetically sealed receptacles are also available. For a copy of Catalog KQ/KR-1, write to the firm.

(Continued on page 115A)

The Highest Sensitivity
and the lowest noise.....

Advanced electronic research by Hitachi technicians has now resulted in the development of a superb frame grid type twin triode (6R-HH8) with excellent high gain and low noise characteristics. As a component of the tuner, the 6R-HH8 ensures an excellent picture with a remarkable degree of definition.

fig. 1 Gain characteristics

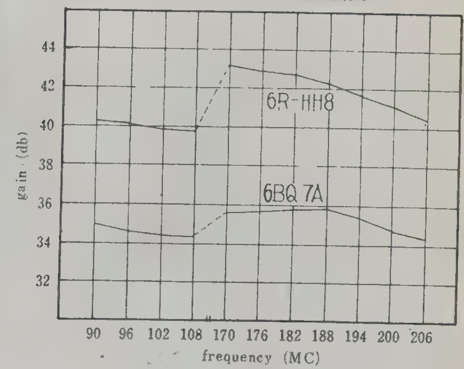
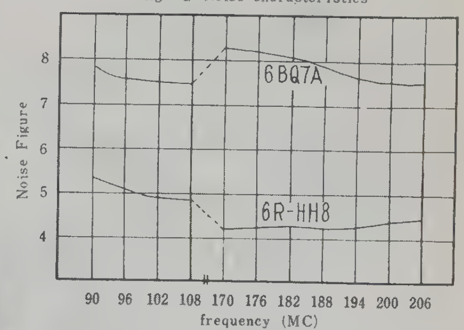
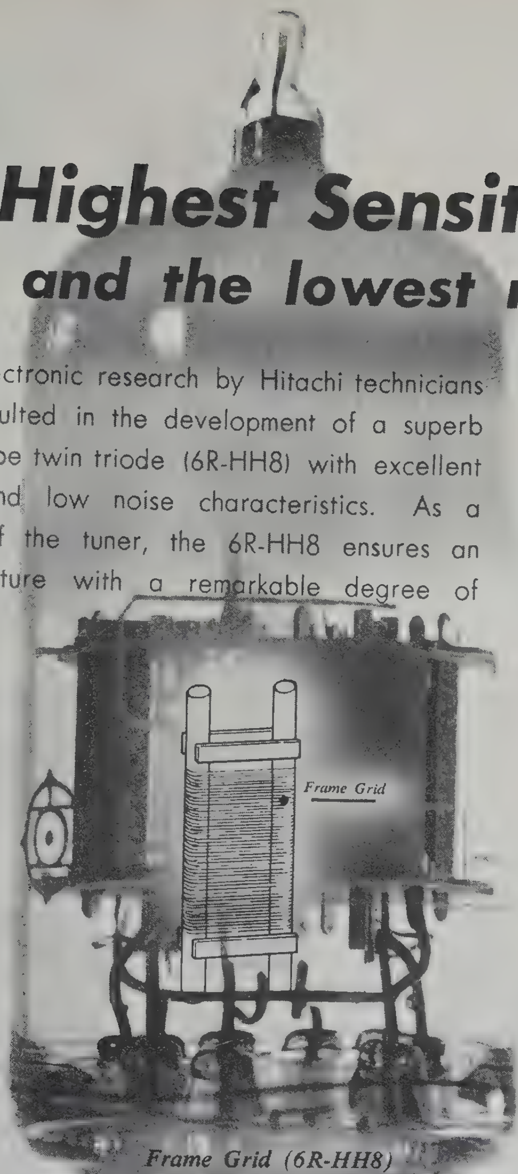


fig. 2 Noise characteristics





Frame Grid (6R-HH8)

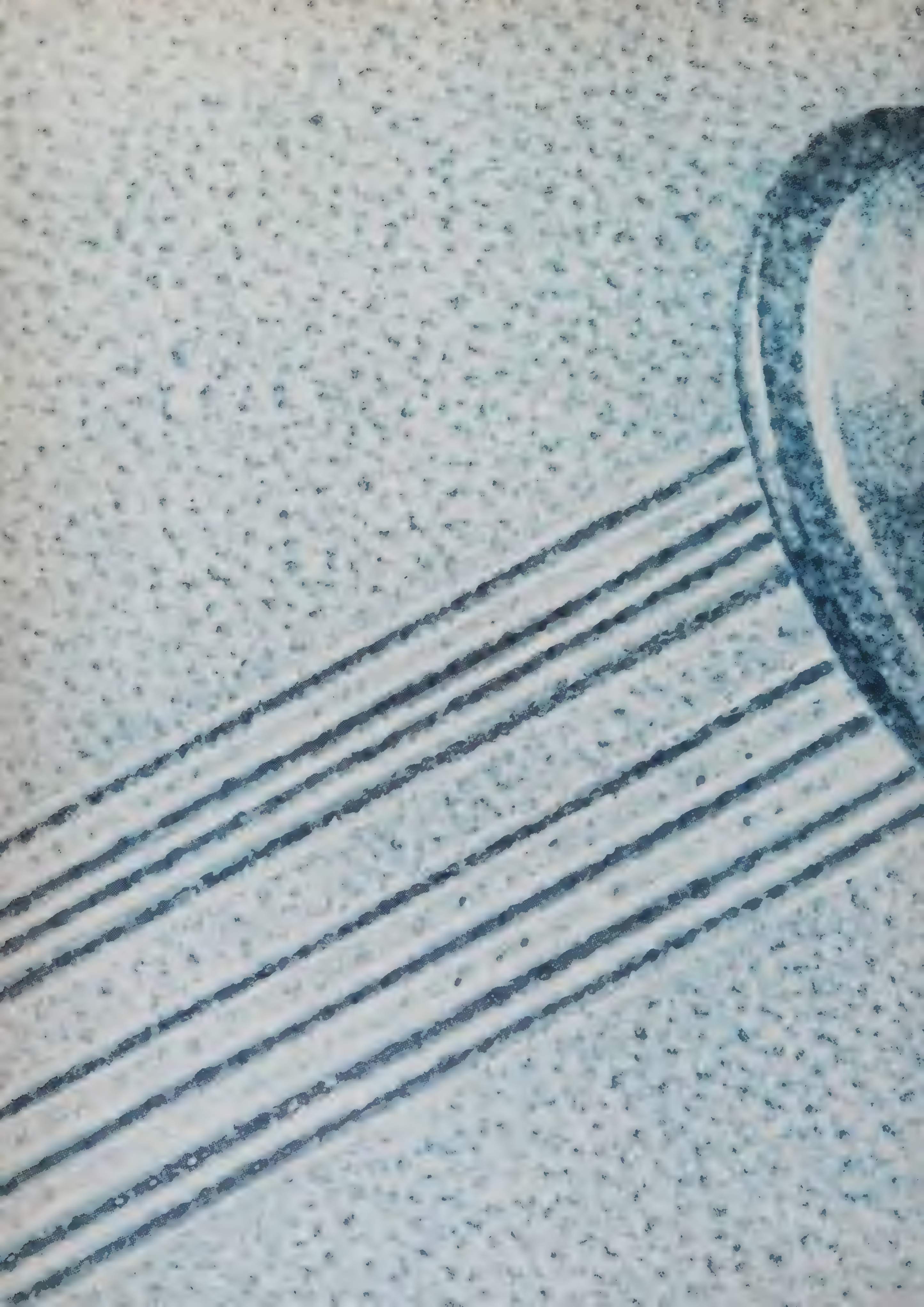
Hitachi also produces other receiving tubes and components for television which, when used together with the new 6R-HH8, cannot fail to earn any maker a market reputation even better than he currently enjoys.

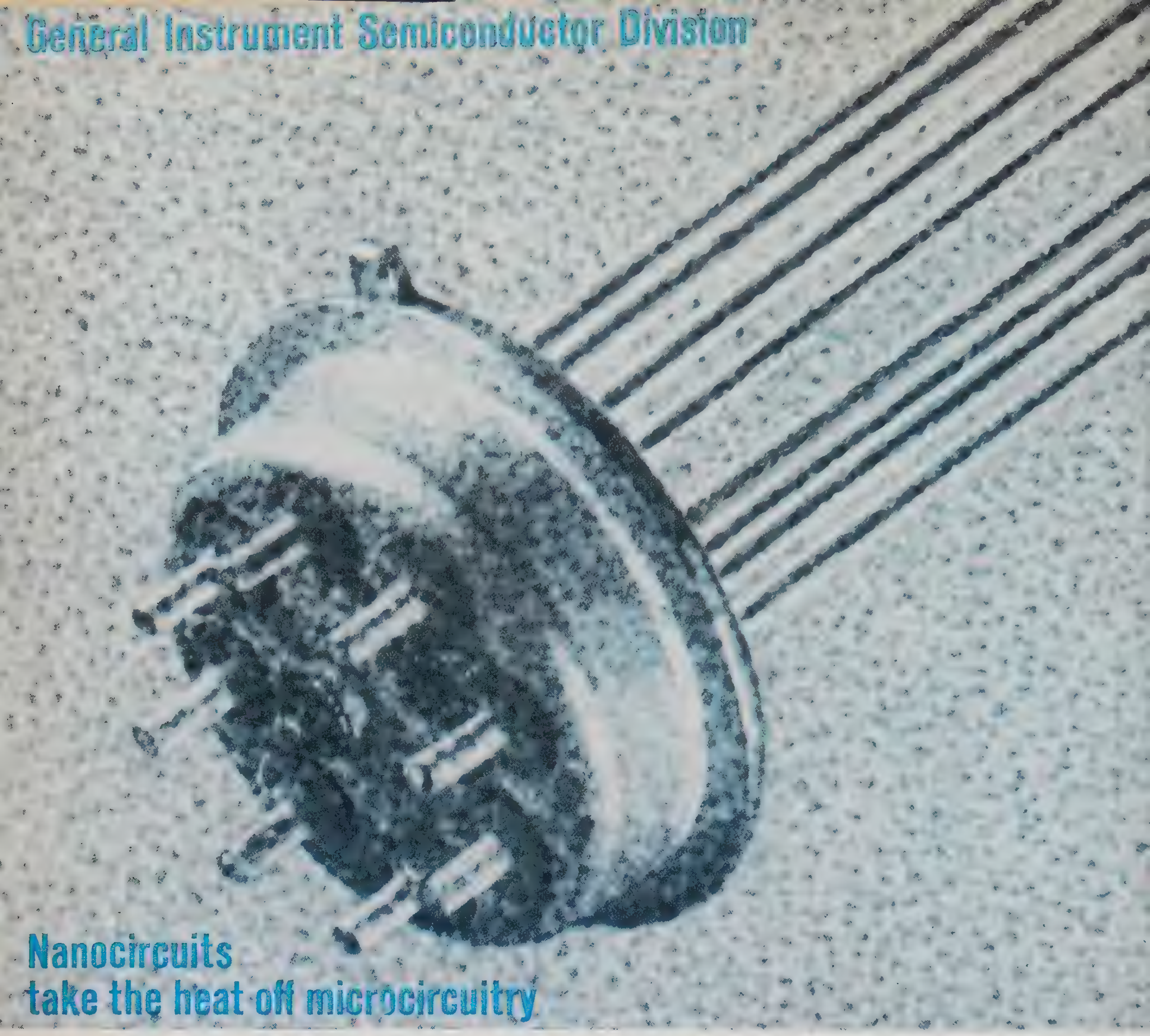
Automatic tube testing equipment

Hitachi, Ltd.

Tokyo Japan

Cable Address: "HITACHY" TOKYO

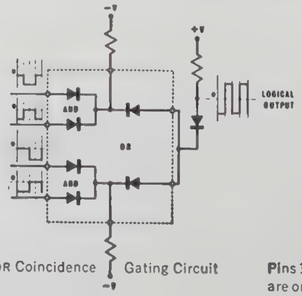




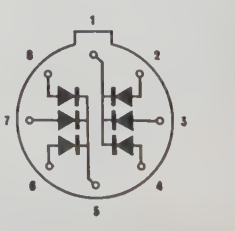
Nanocircuits bring several important advantages to computer logic design, not the least of which is size reduction. This one packs six diodes (it could have been a diode-transistor combination) into a standard TO-5 case. ■ Equally important in the General Instrument concept: only the active components (surface-passivated for stability) are fused to the common substrate. The diodes are not exposed to the heat of such loss-generating components as resistors and capacitors whose demands differ from those of the active elements. ■ Not only is component reliability increased but, since the semiconductors are pre-selected from a 100%-tested standard product line, the designer can evaluate circuit reliability rather than that of individual components. This technique reduces the number of assembly and testing operations, so cost is lower, too. ■ General Instrument also allows the logic designer the flexibility of transferring new or existing circuits, breadboarded with conventional components, directly into nanocircuits. Let us show you how.

Get complete details on nanocircuits and other semiconductor devices from one of our sales offices or the franchised distributor nearest you. Or write today for Bulletin NC-10 to General Instrument, Semiconductor Division, 65 Gouverneur Street, Newark, New Jersey.

GENERAL INSTRUMENT SEMICONDUCTOR DIVISION



AND/OR Coincidence Gating Circuit

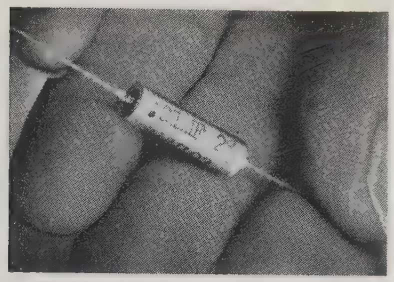


Pins 1 & 5 are 'n' commons. Pins 2, 3 & 4 are one set of 'p' leads; 6, 7 & 8 the other.

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 110A)

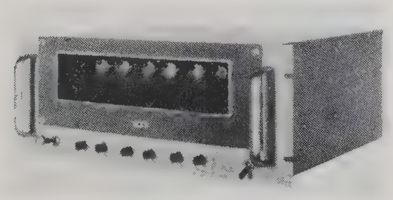
Subminiature Film Capacitors



A new series of subminiature teflon capacitors with low temperature coefficient and high stability is now being manufactured by **Component Research Co., Inc.,** of 3019 S. Orange Drive, Los Angeles 16, Calif. These capacitors have been developed for and designed into critical guidance systems in spacecraft and missile systems. Various temperature coefficients are available from zero to 120 PPM/°C for precise matching with precision resistors and inductors. These precision capacitors have the self-healing characteristics for high reliability of metallized film capacitors but are said to be superior in performance to metallized film and conventional teflon foil capacitors. Tolerances to 0.1% are available. Subminiature metallized mylar capacitors are also available utilizing exceptionally rugged construction and batch uniformity.

Solid-State Time Generator

Model 320 is a new, all-electronic digital time generator for system control and synchronization, or for time correlation among diverse data acquisition and logging devices. Offered by the **Information Technology Div., Lockheed Electronics Co.,** P.O. Box 446, Metuchen, N. J., the instrument employs a 1-kc tuning-fork oscillator to assure time-base accuracy of 5 parts in 100,000 for any clocked interval.



The generator yields two sets of outputs. The basic pulse frequency of the tuning-fork oscillator provides timing markers at 0.001-second intervals and, by means of transistor flip-flop dividers, every 0.01, 0.1, and 1 second as well. Decimal outputs to a total of 23 hours, 59

(Continued on page 116A)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 115A)

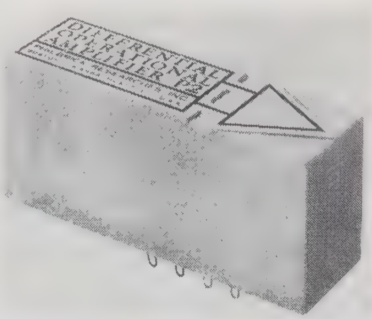
minutes, 59 seconds are also provided at a rear-panel connector, and are simultaneously displayed on an array of six in-line glow-tube decimal indicators (Nixies). An optional extra set of outputs duplicates the decimal values in binary-coded decimal (8-4-2-1 or other) form.

Rack mounting is standard. The clock panel occupies 7 inches vertically, and is 15 inches deep. A removable front plate provides quick access to replaceable, repairable printed-circuit cards. The glow-tube array is contained in a beveled recess for wide-angle viewing and visual contrast.

The basic Model 320 is priced at $2950. Optional configurations provide different clock ranges, BCD outputs, and either greater or lesser time-base accuracy (by means of a crystal-controlled oscillator synchronized to WWV, or by reliance on 60 cps power alone).

Differential Amplifier

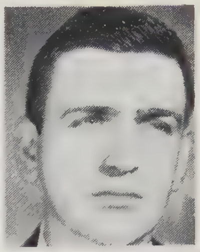
An all-solid state Differential Operational Amplifier, Model P2, has been developed by **George A. Philbrick Researches, Inc.,** 285 Columbus Ave., Boston 16, Mass. Containing neither tubes nor mechanical choppers, it is a true dc differential amplifier whose input is entirely "floating" and yet has a long term drift stability in the sub-millivolt region. Its typical input current is less than a tenth of a millimicroampere, thus making possible its use in long time-constant integrating circuits using small polystyrene capacitors ($R = 10M$, $C = 0.1$ μf for a 1 second integrator), or its use in electrometer type amplifier circuits. There is no limit on the size of a common mode signal save the dielectric strength of the insulating materials.



Its internal dissipation is about 300 milliwatts, fully loaded. Its size is $1\frac{1}{4}$"W $\times 1\frac{11}{16}$"H $\times$ 4"L, and it operates from a ± 15 volt dc supply. Typical open loop gain is over 30,000, frequency response is tailored to give a smooth roll-off with the unity-gain-frequency above 75 kc.

Model P2 is suited to high reliability test and control equipment, although it was designed originally for computing systems. Its sturdy cast case, low temperature rise, and differential input make the P2 quite as much at home in a process control as in a computing system. Because of its low power drain (10 ma), it is easily incorporated into battery operated measuring instruments. Cost is $185 in lots of 25.

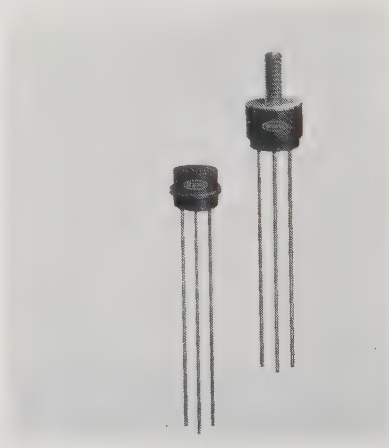
Schlenker VP of PRL Electronics



Robert M. Schlenker has been appointed to the post of vice president, it was announced by **PRL Electronics, Inc.,** 232 Wescott Drive, Rahway, N. J. PRL produces specialized test equipment and power supplies. Schlenker's past experience includes positions with the NJE Corporation as production manager, Douglas Laboratories as a production supervisor, and Federal Tele Communications Laboratories as an engineering assistant. His education includes two years at Union Junior College and two years at the Newark College of Engineering where he studied electronic engineering.

Controlled Rectifiers

Designated as a companion to the 2N1595 Series, **Solid State Products, Inc.,** One Pingree St., Salem, Mass., announces the introduction of their 2N1881–2N1885 Series. This new device group offers improved system performance over wider design limits. Cutoff currents and gate firing sensitivity are improved by a factor of 5. In addition, the new types have higher temperature capability at specific current levels. Complete specifications are available in Bulletin C415-04.



These units will actually replace the 2N1595 Series in many high performance applications, since circuit margins can be improved by orders of magnitude at very nominal cost.

Now in full production, Types 2N1881 through 2N1885 are immediately available in TO-9 Outline, featuring the proven SSPI cold-welded package with all leads isolated, no need for mica washers chassis isolation schemes.

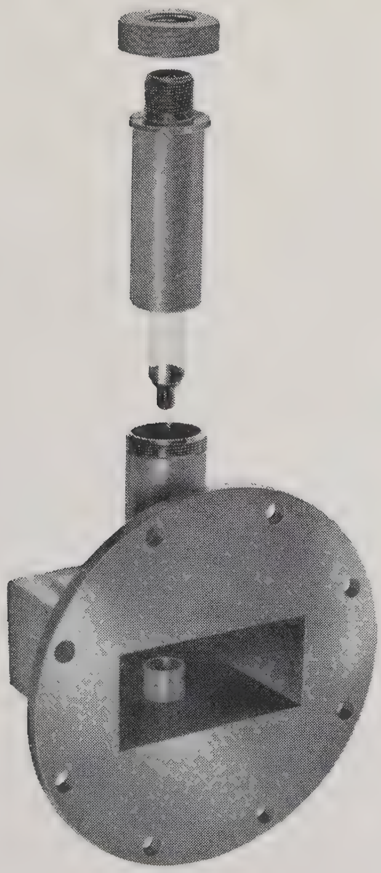
(Continued on page 128A)

For noise at microwave frequencies, too, there's an ideal device in a small package. It's the Litton L-2000 series of miniature gas discharge noise sources. Use them for automatic monitoring of the performance and sensitivity of modern radar systems. They're available to cover the most-used frequency bands and come in a variety of mount configurations.

Acoustical noise: 85 db. at 6 inch distance

The series features a shielded cathode, low modulator drain, and field-replaceable tube insert. Rugged. Insensitive to a wide range of ambient temperatures. Compactly engineered for demanding air and ground environments. Economical because of replaceability, plus added advantages of logistic simplicity and ease of maintenance.

BIG NOISE FROM A SMALL SOURCE



The tube pictured here is the single-ended L-2000 with the LR-2000 insert, specified for a recent generation of FAA airport surveillance radars and for a variety of well-known "S-band" military systems.

For more data on these or other precision gas tube products, write Litton Industries, Electron Tube Division, 960 Industrial Road, San Carlos, California. Or telephone LYtell 1-8411.

GAS NOISE TUBES

Type Number	Frequency Range (megacycles)	Excess Noise Ratio (db)	Nominal Operating Current (ma)	Nominal Operating Voltage (volts)	RF Coupling
L-2008	200-250	18.5 ± 0.5	25	200	¾″ coax*
L-2013	570-630	18.5 ± 0.5	25	200	¾″ coax*
L-2006	1200-1400	18.5 ± 0.5	50	175	¾″ coax**
L-2000(R)	2700-2900	18.5 ± 0.2	75	30	RG-48/U WG*
L-2018(R)	2700-2900	15.5 ± 0.2	75	35	RG-48/U WG*
L-2011(R)	3300-3700	18.4 ± 0.2	150	30	RG-48/U WG*
L-2009(R)	3400-3700	15.5 ± 0.5	125	20	RG-48/U WG*
L-2007	2000-4000	18.5 ± 0.5	85	135	¾″ coax**
L-2010	2000-4000	15.0 ± 0.5	40	60	¾″ coax**
L-2001(R)	5400-5900	13.0 ± 0.5	100	55	RG-49/U WG**
L-2002(R)	7500-8600	14.5 ± 0.5	100	40	RG-51/U WG**
L-2003(R)	8500-9600	14.5 ± 0.5	100	45	RG-52/U WG**
L-2004(R)	8500-9600	18.5 ± 0.5	100	45	RG-52/U WG*
L-2017(R)	8970-9190	18.5 ± 0.5	100	45	RG-52/U WG*
L-2005	16000-17000	18.5 ± 0.5	55	55	RG-91/U WG*

(R) denotes replaceable gas tube insert
*single ended mount
**double ended mount

LITTON INDUSTRIES
Electron Tube Division
MICROWAVE TUBES AND DISPLAY DEVICES

Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

TEACHING POSITIONS

PROFESSORS all levels. Ph.D. required. Solidstate electromagnetics, and engineering analysis especially. Income with research, competitive with industry. Graduate program as large as undergraduate. Mile high, dry climate.

INSTRUCTORS, Associates, Assistants for teaching and research part-time positions while working for M.S. and S.Sc. Large graduate program assures variety of offerings. Apply Chairman, E.E. Dept., University of New Mexico, Albuquerque, New Mexico.

BROADCAST ENGINEER

Long term European assignment with private organization engaged in construction and operation of large high-powered shortwave radio stations. Top qualifications, experience and executive ability desired. Good salary plus overseas benefits. Inquiries treated in confidence. Reply to Box 2043.

ASSOCIATE PROFESSOR

Electrical Engineering faculty being expanded in relatively new and rapidly growing department, with positions available to rank of Associate Professor with initial salary range to $6000 for base year of 9 months, depending upon education and experience. Further opportunity for research and other programs in this industrial area. Preferred background emphasis in field and circuit theory, electronic systems and control. Address full background to Chairman, E.E., University of Bridgeport, Bridgeport 4, Conn.

ELECTRONICS SECTION HEAD

Cornell Aeronautical Lab. has an attractive opening as technical supervisor of a section involved in the solution of complex problems associated with the electronics field, especially in the areas of weapons systems, reconnaissance and surveillance, and guidance and control. Position holds outstanding opportunities for growth from an already well established research program, both by way of personnel and highly successful technical projects. Horizons not limited by marketing and sales. The Laboratory provides a desirable combination of the best features of industry as to financial and equipment resources coupled with a university associated atmosphere for constructive R&D activity at the highest level. Salary range starts at $13,500 depending on qualifications and experience. Send complete resume to F. L. Rentschler, Cornell Aeronautical Lab., 4455 Genesee St., Buffalo 21, N.Y.

ELECTRONIC ENGINEER

Electronic engineer to teach lecture and laboratory courses. Up-to-date knowledge of the field required. Working and living conditions excellent: salary and opportunity good. Write to Dean of Engineering, California State Polytechnic College, San Luis Obispo, Calif.

(Continued on page 123A)

THANKS FOR SHARING THE LOAD, DR. MAXWELL!

Your equations together with Newton's Laws serve as a basis for explaining classical electromagnetic phenomena. Most important among the outgrowths of your theory are radio and its allied invention, radar. At AC, we are using techniques for the generation and propagation of electromagnetic waves to increase the total capabilities of the B-52 weapons system.

If you are interested in applying yesterday's theories, like Maxwell's, to today's Mach 2 and 3 aircraft, and if you have a BS, MS or PhD in EE, ME, Physics or Math, please contact Mr. G. F. Raasch, Director of Scientific and Professional Employment, Dept. B, 7929 South Howell, Milwaukee 1, Wisconsin.

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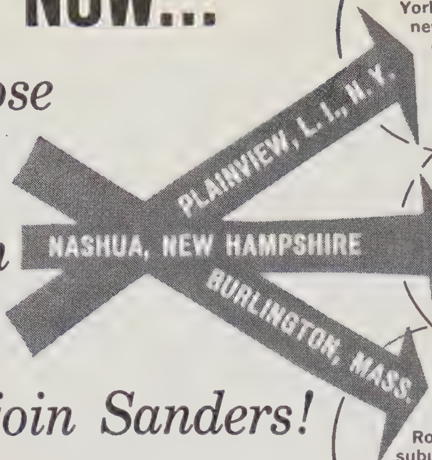

suburban New York City. Brand new facility opened in December 1960.

Company headquarters in the beautiful hill country only an hour from down- town Boston.

on famous "Electronics Row" in suburban Boston. Advanced Systems Laboratories opened in November 1960

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STARTING only 9 years ago with 11 Engineers and an initial order of $117.00, Sanders Associates today has a personnel role of over 1600—and a contract backlog of $53,000,000.

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Pioneering programs are being continued in phased arrays, radar, pulse doppler radar systems, space radar and communication systems, providing stimulating assignments in space technology, missiles and radar systems.

To arrange a convenient interview appointment, send resume in confidence to R. W. McCarthy.

POSITION IN NASHUA

SENIOR CONSULTANT TRANSISTOR CIRCUITRY

To provide technical guidance at the Corporate level on a wide variety of transistor circuit design problems. Requires ability to design detailed circuits rapidly.

POSITIONS AVAILABLE AT ALL LOCATIONS FOR:

SENIOR SYSTEMS ENGINEERS

To contribute to advanced techniques in the general field of military electronic systems. Applicable experience includes systems analysis, synthesis and integration, with extensive background in circuit design augmented by hardware implementation.

CIRCUIT DESIGN ENGINEERS

EE or Physics graduates with 2 to 8 years experience and familiarity with tubes and transistors and their utilization in all types of circuits, as well as the integration of circuits into subsystems.

TRANSMITTER DESIGN ENGINEERS

2 to 8 years experience. For work up to and including microwaves.

PRODUCT DESIGN ENGINEERS

ME with heavy experience in feasibility studies coupled with experience in taking developed systems into production, monitoring mechanical design and overall packaging concepts of ECM or other airborne systems.

POSITIONS IN PLAINVIEW, LONG ISLAND

GROUND SUPPORT EQUIPMENT ENGINEERS

To design and develop system, assembly and sub-assembly electronic test equipment for the military. Should have appreciation for test equipment philosophy, with extensive experience in circuit design and hardware follow-through.

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SANDERS ASSOCIATES, INC.

NASHUA, NEW HAMPSHIRE

Major Expansion in the program of the Laboratory requires participation of senior members of the scientific community in our programs:

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SYSTEMS:
- Space Surveillance
- Strategic Communications
- Integrated Data Networks

NEW RADAR TECHNIQUES

SYSTEM ANALYSIS

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- Psychology
- Theory

INFORMATION PROCESSING

SOLID STATE Physics, Chemistry, and Metallurgy

• A more complete description of the Laboratory's work will be sent to you upon request.

Research and Development

LINCOLN LABORATORY

Massachusetts Institute of Technology

BOX 16

LEXINGTON 73, MASSACHUSETTS

Positions Open

(Continued from page 118A)

ASSOCIATE DIRECTOR AND DIRECTOR

Associate Professor and Director of Research in Electrical Engineering. To have the responsibility for developing a research program including the securing of research contracts and the directing of research activity. Salary depending on experience and academic background. Write to Head, E.E. Dept., South Dakota School of Mines and Technology, Rapid City, South Dakota.

TWO ELECTRONIC ENGINEERS

Good theoretical background, one with at least 7 years' experience, and the other with at least 2 years' in research and/or development. Excellent career possibilities in rapidly expanding industry. Congenial atmosphere. Call Personnel Dept. Riverview 9-4800. Massa Div., Cohu Electronics, Inc., 280 Lincoln St., Hingham, Mass.

PROFESSOR

The University of Alaska has an opening for an Assistant Professor of Electrical Engineering to teach and to do research on the ionosphere, the aurora, or on Northern problems in communications or power. Industrial experience or Masters' degree required, or for an instructor with less preparation. This is a pioneer university in a new state. We work and live in conventional surroundings, but beyond us is wilderness. Write Air Mail to Dept. of E.E., University of Alaska, Box 497, College, Alaska.

COMMUNICATION ENGINEER

College graduate with several years' experience and good technical background, including radio system application or installation. Consulting engineering firm in New York City. Box 2044.

ASSOCIATE PROFESSOR OF E. E.

Ph.D. required and special competence in electromagnetic theory desirable. Should have some teaching and research experience. Position available Sept., 1961. Write Head of E.E. Dept., State University of Iowa, Iowa City, Iowa.

ELECTRONICS ENGINEER

Inductive Devices. Well established and growing company located in Culver City, Calif., has an excellent opportunity for a capable engineer to organize and manage the design and manufacture of low pass, high pass and band pass filters and electro magnetic delay lines. Salary will be commensurate with training, experience and ability to manage. Liberal company benefits. Box 2047.

ELECTRICAL ENGINEERING TEACHING POSITIONS

Ph.D. degree required. Teaching experience desirable but not necessary. Excellent opportunity for young man interested in teaching electronics, network theory, control systems and computers at undergraduate and graduate level. Appointment effective Sept. 1961. Write, Chairman, E.E. Dept., University of Houston, Houston 4, Texas.

ASSISTANT & ASSOCIATE PROFESSOR

Applications are invited for Assistant and Associate Professor of E.E. Candidates should be well qualified academically, preferably to the doctorate level, and should have some research, design or teaching experience in control systems. Duties include teaching at undergraduate and graduate levels, organization and direction of laboratory classes, conducting research and super-

(Continued on page 125A)

Consider, for example, basic diffusion studies of P impurities in N germanium, and associated problems in thermal conversion and impurity density distribution./That's where Morris Chang (MIT '52) started as product engineer at Texas Instruments Semiconductor-Components division in 1958. His immediate objective was to develop NPN diffused base HF germanium transistors. But more than that, his overriding goal was *to develop a producible product.* With impressive speed, Chang moved his product from original idea into mass production and saw it successfully sold to military and commercial markets./Chang, too, moved forward. He was made supervisor of diffused device engineering and today has advanced to manager of device development in TI's germanium products group, working on very HF switchers and amplifiers, epitaxial techniques and surface and reliability studies./Unique success story? No — an example of the professional growth opportunity fostered by a fast-growing, aggressive company. /Unique opportunity? We think so. TI is engineer-managed, sales-minded. It all adds up to a rewarding environment for the engineer who wants to see his ideas take form as new products.

SEMICONDUCTOR-COMPONENTS DIVISION

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RSVP We'd like you to know more about TI: its opportunities and the rewarding way of life that its Texas location affords. We'd like, at the same time, to know more about you. A confidential exchange of information may prove the catalyst to lead you to the professional fulfillment you've established as your personal goal. Write: C. A. Besio, Dept. 113, Texas Instruments Incorporated, Semiconductor-Components Division, P. O. Box 5012, Dallas 22, Texas.

Positions Open

(Continued from page 123A)

vising research students. Salary scales are open and competitive with those of industrial and research establishments. Additional stipends are offered to professors who remain on the campus for 11 months of the year and carry out research during this period. Write to Chairman, Dept. of E.E., McMaster University, Hamilton, Ontario.

DESIGN ENGINEER

Design Engineer for low noise and wide band UHF and VHF amplifiers. Outstanding opportunity. Chief Engineer potential. Small, vigorous firm, located in beautiful central Pennsylvania. Educational opportunities across the street at Penn State. Stock option. Send resume to Community Engineering Corp., P.O. Box 824, State College, Pa.

TEACHING POSITIONS

The E.E. Dept. of the City College of New York has several positions available on the teaching staff beginning Sept. 1961. Rank and salary commensurate with qualifications and experience. Opportunity for graduate study. Applicants must be present residents of the U.S. Address inquiry to Prof. H. Taub, Dept. of E.E., The City College, Convent Ave. at 139th St., New York 31, N.Y.

SENIOR ELECTRONIC ENGINEERS

Electronic Engineers are needed for development of new types of Power Supplies and other electronic instruments. Experience is desired in the fields of power supplies, AC line regulators, electronic instruments and magnetic amplifier and transistorized circuits. Salary is open and is commensurate with applicant's background and ability. Company benefits. Apply Perkin Electronics Corp., 345 Kansas St., El Segundo, Calif.

ENGINEERS

Openings for Electrical Engineering Dept. Chairman and for Assistant or Associate Professor. Must have Doctorate. Opportunity for research in Bio-medical Electronics. Excellent salary and environment. Send resume to Chairman, Dept. of E.E., University of Vermont, Burlington, Vermont.

SCIENTIST—TRANSLATORS RUSSIAN TO ENGLISH

Supplement your income while you keep abreast of latest Soviet research in your field. Guaranteed steady flow of translation in your specialty. Must have graduate study and experience in Physics (atomic, plasma, solid state) Crystallography, Instrumentation, Automation. Native command of English necessary. Chinese-English scientist-translators also wanted. Apply Consultants Bureau, 227 West 17th St., New York 11, N.Y.

ASSISTANT OR ASSOCIATE PROFESSOR

Assistant or Associate Professor of Electrical Engineering required to teach graduate courses and to do research. Ph.D. required, teaching experience desired. Expanding department with large graduate program. Write to Dr. F. B. Haynes, Dept. of Electrical Engineering, Drexel Institute of Technology, Philadelphia 4, Pa.

TEACHING—ELECTRICAL ENGINEERING

Applications are invited for positions in the Electrical Engineering Dept. in the fields of circuit analysis, digital computers, electrical measurements, communications, control systems, and industrial electronics. Rank and salary are dependent upon qualifications. New engineering building and equipment are to be in use by commencement of the fall term. Applications should be addressed to the Acting Chairman, Dept. of E.E., Essex College, Windsor, Ontario, Canada, from whom further particulars may be obtained.

ASSISTANT CHIEF ENGINEER

Electronics engineer with management experience interested in assisting in direction of commerical television operation in major mid-western community. To direct and improve operation of large department of engineers and technicians. Must be well organized, conscientious, management-oriented. Salary to $12,000. Box 2050.

TWO ASSOCIATE PROFESSORSHIPS

$7,900-$10,300 in Electrical and Mechanical Technology. Advanced degrees and industrial experience desirable. These are teaching positions with good opportunities for those interested in administrative work. Write to Electrical-Mechanical Technology Dept., Bronx Community College, 120 East 184th St., Bronx 68, N.Y.

COMPONENTS ENGINEER

BSEE. degree. Salary open. 4 years' experience in application of component parts in military electronics equipment. Familiar with MIL specs applicable to airborne and missile electronic equipment. Will specify and evaluate resistors, capacitors, magnetics, semi-conductors, connectors, relays and switches. Computer Systems Laboratory. Send resume to Don Colvin, Litton Systems, Inc., 5500 Canoga Ave., Woodland Hills, Calif.

Positions Wanted

By Armed Forces Veterans

In order to give a reasonably equal opportunity to all applicants and to avoid overcrowding of the corresponding column, the following rules have been adopted:

The IRE publishes free of charge notices of positions wanted by IRE members who are now in the Service or have received an honorable discharge. Such notices should not have more than five lines. They may be inserted only after a lapse of one month or more following a previous insertion and the maximum number of insertions is three per year. The IRE necessarily reserves the right to decline any announcement without assignment of reason.

Address replies to box number indicated, c/o IRE, 1 East 79th St., New York 21, N.Y.

ENGINEERING TECHNICIAN

6 years experience in Military communications systems, and miniature audio systems. Will graduate from RCA Institutes (advanced electronics course) in February of 1961. Some college background. Age 28; married. Box 3016 W.

ELECTRICAL ENGINEER

Seeking management position. Nearly 10 years of successful experience in planning, organizing, and administering of various phases of projects in the development of electro-mechanical products from specifications through prototype and final test. Box 3017 W.

(Continued on page 126A)

Positions Wanted

By Armed Forces Veterans

(Continued from page 125A)

CONSULTING ENGINEER

BSEE. 1958, EIT certificate 1958. 2 years technical publications experience with technical and managerial training at the graduate level. 1 year Naval electronic school and 1 year as a Naval electronics technician. Other electrical utility and heavy steel mill control experience. Member IRE, Associate Member AIEE. Age 27. Desires association with an architectural and engineering firm. Box 3018 W.

TRANSFORMER DESIGNER

9 years experience in design, production, test and tooling for a wide variety of small transformers. Particular emphasis on design for epoxy encapsulation and on design of Hi Q reactors and low level multi-shielded units. Box 3019 W.

ENGINEER

Age 33; married; 1 child. Navy Electronic Technician. BEE. Analytical, imaginative, artistic; not oratorical. 10 years experience with airborne electronic equipment. Seeks firm in Oklahoma, Minnesota, upper New York or Oregon which can express logical product-demand evaluation. Box 3028 W.

ENGINEER

Interested in production. B.S. in E.E., MBA. in Production. Completed Active Duty. Box 3929 W.

EDITOR—PUBLICATIONS MANAGER

B.S. in Physics; 10 years technical writing experience, including 4 years as supervisor; 3 years teaching radio and television repair, laboratory and theory; holder of 1st class radiotelephone license; excellent mathematician; working knowledge of French, German and Russian; expert typist and stenographer. Desires position as editor of electronics publication or as supervisor in publications section of electronic equipment manufacturer. New York City area preferred. Box 3930 W.

PATENT ADMINISTRATION

Age 40. Engineering and Law degrees. 9 years experience prosecuting and preparing patents in the complex electronics arts in radio, radar, TV, scopes, UHV, VHF fields. Desires position in small company or law firm anywhere in U.S. Box 3931 W.

ENGINEER—PILOT

Age 27; BSEE., 1961; 6 years experience piloting USAF all-weather interceptors; presently Assistant Operations Officer, Instrument Flight Examiner, and Instructor Pilot in F-102A and T-33A Aircraft of an ADC Squadron; 1500 Jet hours. Desires position as aircrew member testing airborne electronic equipment. Box 3932 W.

DIGITAL COMPUTER ENGINEER

MEE. Experienced in all phases of digital computer engineering, including significant supervisory, circuit design, logical design, system development. Extensive magnetic core experience. 12 years experience, including sales, executive, college teaching. Patents. Desires stimulating, forward looking work. Box 3933 W.

(Continued on page 128A)

Positions Wanted By Armed Forces Veterans

(Continued from page 126A)

MANAGER, GRADUATE ENGINEER AND PILOT

10 years experience in weapon and automation systems R & D. Desires managerial position in small, technically oriented company. Available in June. Resume upon request. Box 3934 W.

ELECTRONICS ENGINEER

BSEE.; age 30. Desires position in electronic systems engineering, marketing and management. 6 years experience in radar, communications, television and missile electronics. Foreign assignments preferred. Box 3940 W.

ELECTRONICS TECHNICIAN

Graduate RCA Institutes V7 course. Age 28, married. Experience in repair of spectrophotometers, pH meters, hemophotometers, titrimeters, G. C., etc. Knowledge of Danish, German, French and Spanish. Seeks position with bio-medical manufacturer as technician or representative. Willing to relocate to Denmark, Sweden or in U.S. Box 3941 W.

ELECTRONICS TECHNICAL WRITER (FREE LANCE)

Desires free-lance technical writing assignments for short term contracts. Prefer Southern California area. Age 30. 10 years experience in military and civilian electronics. 2 years experience in technical writing. Active member in IRE and STWP. Resume upon request. Box 3942 W.

ENGINEERING MANAGER

Age 36. BSEE., MSEE. Completed course credits for Ph.D.EE. Control, communications and systems engineering. Teaching experience and business training. Presently manager of small consultant team to U. S. Army. Will consider overseas assignment. Box 3943 W.

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

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Power Isoformer

The Model L131 Power Isoformer, manufactured by **Elcor, Inc.**, 1225 W. Broad St., Falls Church, Va., is a 350-watt isolation transformer featuring low distributed capacitance between the secondary winding and ground. Intended for general purpose laboratory use or for mounting as a component in equipment, this transformer permits isolating from ground (i.e. "floating") a complete instrument or a component circuit. So effective is the isolation from the power line and ground that signals from dc to several megacycles can be applied between the chassis of a typical conventional instrument and ground in applications where such operation is necessary or desirable because of interconnection with other equipment.

(Continued on page 131A)

REDEYE ...a *NEW* concept in antiaircraft philosophy, being developed expressly for infantrymen

CONVAIR/POMONA is the developer of REDEYE, a one-man, shoulder-launched guided missile . . . a new lightweight weapon that will strike swiftly with pin-point accuracy . . . a weapon that will ensure our ground forces of the Army and Marine Corps an effective means of combatting low-flying enemy aircraft.

REDEYE is one of several unique and advanced weapon systems being developed at Convair/Pomona. If you have a sincere desire to be a part of this and other important programs, and possess proven capabilities in research, development and production design . . . if you would enjoy living in Pomona Valley, one of Southern California's finest educational, cultural and recreational centers, then complete the attached inquiry.

CONVAIR/POMONA CONVAIR DIVISION OF GENERAL DYNAMICS

NEWS New Products

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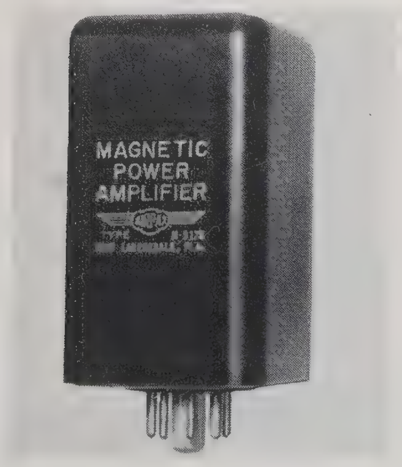
(Continued from page 128A)

Shunt capacitance has been reduced to 40 $\mu\mu$f in this transformer by providing an air gap between the secondary winding and the core. A close fitting electrostatic shield essentially encloses the secondary winding, and the air gap is employed between this shield and the core. A separate shield encloses the primary winding. Excellent magnetic coupling is maintained in order to achieve high efficiency, while electrostatic coupling and distributed capacitance are kept to a minimum. Additional features include a leakage resistance value for the secondary winding in excess of 100 kilomegohms, and a breakdown voltage exceeding 5000 volts.

The Power Isoformer may be employed either for isolating a complete instrument or for isolating a component circuit.

Magnetic Servo Amplifiers

These amplifiers were designed by **Airpax Electronics, Inc.,** Cambridge, Md., to drive instrument type two-phase servo motors, controlling the current to both phases, thereby greatly reducing the stand-by power drawn by equipments employing them.



Input is dc polarity reversible; output is ac phase reversible. The combined power gain of this power amplifier used with a

(Continued on page 132A)

NEWS New Products

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(Continued from page 131A)

pre-amplifier is 8×10^6. As a result, the input may be as low as 75 microamperes for maximum output to the motor.

Delivery is three to four weeks. Priced at $93.00 each in quantities 1–6. For further information contact the company direct.

Voltage Calibrator

A new laboratory voltage calibrator is now available from **Dynage, Inc.**, 75 Laurel St., Hartford, Conn., to fill the long

(Continued on page 134A)

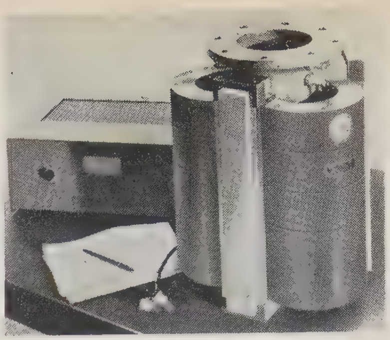
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 132A)

awaited needs of the laboratories and process industries for fast, convenient and accurate calibration of self-balancing potentiometers, oscillographs and all other instruments having a high input impedance. The calibrator operates from 115 vac ($\pm 20\%$). It is packaged in an attractive mahogany box with a black formica cover. Standard ranges are 0 to ± 1000 mv, 0 to ± 100 mv, 0 to ± 10 mv dc. The voltage stability is $\pm 0.01\%$ or less for line voltage variations of 94 to 140 volts. The temperature stability is $\pm 0.002\%/°$ C over a range of 0° C to 40° C. The linearity is $\pm 0.05\%$. The overall accuracy using the sum of the worst stability conditions above is 0.1% of full scale. The size of the unit is $8\frac{1}{2}'' \times 8\frac{1}{2}'' \times 4\frac{1}{2}''$. Battery (internal) powered portable units and rack mounted units are also available.

Ion Pump

The availability of a new, large vacuum ion pump for use in the ultra-high vacuum field has been announced by **Consolidated Vacuum Corp.**, 1775 Mt. Read Blvd., Rochester 3, N. Y., a subsidiary of Consolidated Electrodynamics/Bell & Howell.



The new PDV-300 DriVac pump is an electronic getter ion pump utilizing a three-electrode construction which is said to insure higher pumping speeds and eliminate pressure surges when operating against inert gases.

The features of the PDV-300 DriVac pump make it especially useful for ultra-high vacuum applications such as semiconductor processing, vacuum tube processing, electron microscopes, particle accelerators, and field emission, electron beam and molecular beam studies. It has a nominal pumping speed of 300 liters/second with an ultimate pressure of 2×10^{-10} mm Hg—the lowest reading obtainable on a commercial ionization gage as determined by the X-ray limit of the tube.

A stainless steel pump casing, capable of withstanding bakeout temperatures up to 450° C, has a standard 6-inch 150-pound ASA flange with a flat gasket surface suitable for a metal wire gasket seal. The only

(Continued on page 136A)

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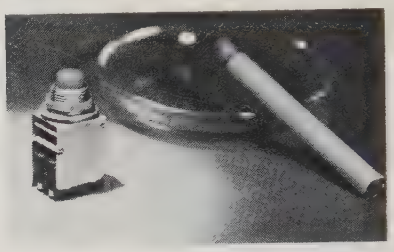
part of the new DriVac pump that is subject to wear is the sputter-cathode which has an expected life of 200,000 hours if the pump were operated at a set pressure of 1×10^{-7} mm Hg. However, since the expected life of the sputter-cathode is inversely proportional to pressure, even longer life can be expected at lower pressures.

Complete information is available in Bulletin 6-2 from the firm.

Snap-Action Switch

A new miniature push-button switch designated B-PB3-2, for panel mounting, has been developed by the **Milli-Switch Corp.**, Gladwyne, Pa. It conforms to MIL-S-6743, Military Part 25085-1.



(Continued on page 138A)

Honeywell Aero... for the best of both

CLIMATE FOR CREATIVITY—Honeywell Aero is now producing inertial platforms for the Polaris Missile. In addition Aero Division Engineers have created an Electrically Suspended Gyro for use on Polaris launching submarines, which is capable of providing accuracies never before achieved in an inertial navigation system. This project is typical of the creative concepts and ideas which are being evolved and further developed into working hardware at Honeywell Aero. We invite you to share in this creative atmosphere where there is ample opportunity for a man of imagination, drive and talents to grow in professional stature and have his accomplishments recognized and rewarded.

CLIMATE FOR ENJOYMENT—Spectator or participant, you'll find whatever sport interests you in the Minneapolis area. For example, this is the heart of America's finest fresh water fishing country. At the end of a busy day or week you can angle for pike, bass, trout, or scrappy panfish in the more than 80 lakes within 25 miles of Minneapolis. Fishing is just one aspect of the many recreational, educational, social, and cultural pleasures you and your family will enjoy when you work at Honeywell's Aeronautical Division in Minneapolis. For information on specific openings, write: Mr. James H. Burg, Technical Director, Aeronautical Division, 2634 Ridgway Road, Minneapolis 40, Minn.

Honeywell

H Military Products Group

To explore professional opportunities in other Honeywell operations, coast to coast, send your application in confidence to: Mr. H. T. Eckstrom, Honeywell, Minneapolis 8, Minn.

NEWS New Products

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The switch weighs 6 ounces and requires an operating force of 21 ounces. Its electrical rating is: 5.0 amperes at 120/240 volts ac, 3 amperes at 30 volts dc inductive at sea level, 4 amperes at 30 volts dc resistive at sea level.

The switch is also available with rating of: 10 amperes at 30 volts dc resistive at sea level, 8 amperes at 30 volts dc inductive at sea level, 10 amperes at 125 volts ac, 5 amperes at 240 volts ac.

Portable Temperature Chamber

A portable temperature chamber marketed by **Delta Design, Inc.,** 7460 Girard Ave., La Jolla, Calif., has been introduced as Model 1060L. This unit is heated by electric elements and cooled either by liquid CO_2 or dry ice. To cool with liquid CO_2, the user connects the chamber with fittings provided to a syphon bottle of liquid CO_2, then dials his selected temperature. A built-in solenoid-operated valve controls CO_2 flow automatically. On other occasions the user can insert dry ice into a special compartment and cool by this method. Because of the compact size both heating and cooling are fast.

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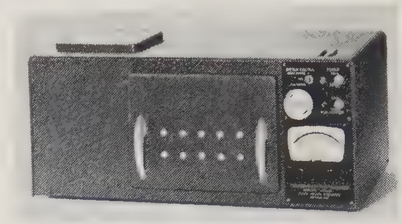
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 138A)



A sensitive meter-relay and thermocouple control temperature within $\pm\frac{1}{2}°$ F anywhere in the range of $-100°$ F to $+500°$ F. Overshoot or undershoot when approaching a temperature setting is eliminated by the control circuitry. A centrifugal blower assures temperature uniformity throughout the test space.

Test volume is $10'' \times 7'' \times 7''$, accessible through a convenient drawer containing 10 sleeves for test leads. The unit operates on 117 volts ac and can be programmed with an auxiliary timer to alternate between a high and a low temperature. The chamber is constructed of stainless steel interior, aluminum exterior with gray wrinkle enamel finish, and rigid glass fiber insulation.

For further information write to the firm.

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NEWS New Products

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Playback Assembly

The PA-8414 magnetic head playback assembly produced by **Applied Magnetics Corp.,** P.O. Box 368, 1407 Norman Firestone Rd., Santa Barbara Airport, Goleta, Calif., is designed to operate with built-in time delay, directly on an FR-100 tape transport. Other precision mounting arrangements are available. There are 14 channels with 0.025 tracks on 0.070 centers

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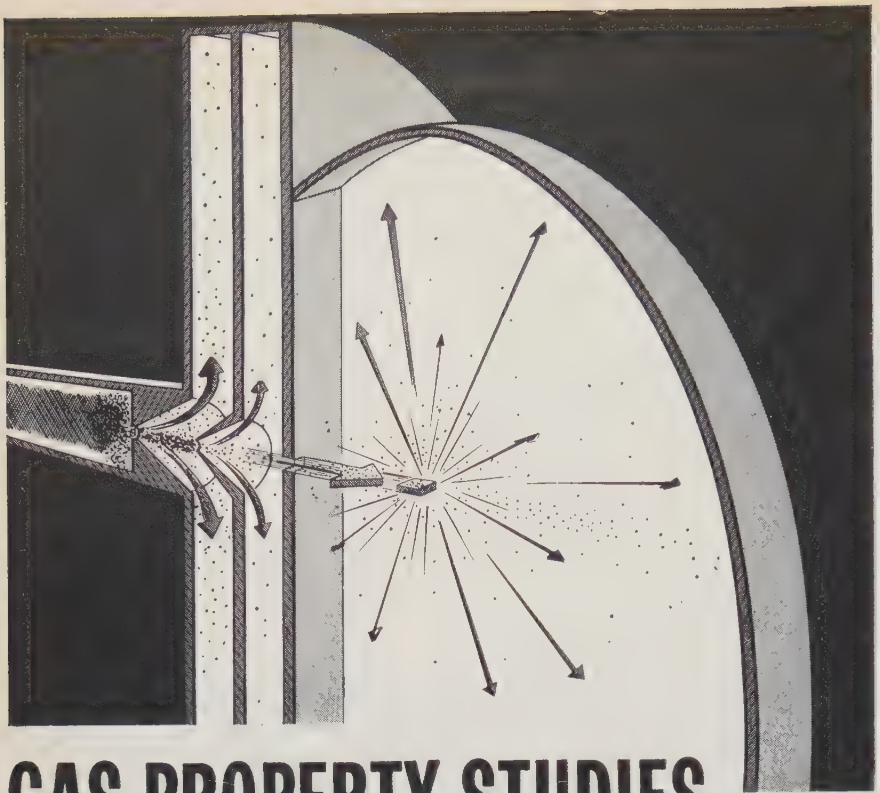

NEWS New Products

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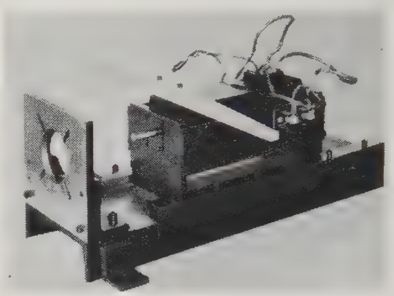
constructed with one movable head micrometer controlled to three divisions per 0.001 movement, and a pre-set time difference introduced between each additional head. This staggered array of heads providing fixed time delays between tracks should find considerable usage in cross correlation work. The same unit is also available having each head individually movable with screw driver adjustment.

Dynamometer

Availability of a new high RPM dynamometer, Model No. D-1101, for drive motor testing, is announced by the **EEMCO Div., Electronic Specialty Co.,** 4612 W. Jefferson Blvd., Los Angeles **16,** Calif.



An example of the capabilities of this new device is found in connection with motor testing of the Minute Man missile. EEMCO has been engaged in the development of motors for a motor pump combination for three stages of this missile, the first stage motor requiring testing at an RPM up to 20,000 with an output of 7 horsepower. The Model No. D-1101 met the requirements.

The D-1101 consists of a generator constructed from an EEMCO motor frame mounted in ball bearings to allow free movement in the horizontal axis. Across the top of the generator is a 20" arm, the end of which is attached to a common weight scale which indicates inch/pounds generated by the test motor.

The test motor, which can be electrical, pneumatic, or hydraulic, is mounted on a Universal mounting plate at the end of the dynamometer frame. A gear box may be installed, if desired, between the motor and the generator for slower speeds. An electronic tachometer may be used to measure the RPM from the driving motor, and this used in conjunction with the inch/pounds indicated on the scale, allows the meter horsepower to be calculated. Output torque is changed by varying the resistance of the rheostat grids connected to the armature of the generator, and also by varying the voltage input to the generator field. A power supply capable of 50 volts, 20 amperes, and a rheostat with a rating up to 4000 watts is required.

(Continued on page 148A)

electronic and electromechanical engineers in a unique role

The engineers and scientists of Aerospace Corporation are in the forefront of a rapidly advancing state-of-the-art in sensing and information systems. Their unique role: a critical civilian link uniting government and the scientific-industrial team responsible for development of space systems and advanced ballistic missiles. In providing scientific and technical leadership to every element of this team, they are engaged in a broad spectrum of activities, from formulation of new concepts to technical review and supervision of hardware development by industry. Specific areas of interest include inertial and radio guidance, automatic control, communications, instrumentation, space- and ground-based computing, telemetering, tracking, auxiliary power, infrared, television, optics, and photography. Now more men of superior ability are needed; highly motivated engineers and scientists with demonstrated achievement, maturity, and judgment, beyond the norm. Such men are urged to write Mr. George Herndon, Aerospace Corporation, Room 109, P. O. Box 95081, Los Angeles 45, California.

Organized in the public interest and dedicated to providing objective leadership in the advancement and application of space science and technology for the United States Government.

AEROSPACE CORPORATION

has a light in his eye that money can't

factor. But a man's real warmth for his job—his real measure

check. To find himself in the right emotional

acknowledged masters in their fields—to be an active

—these are the satisfying, the lasting, rewards. And

talent to come to Lockheed Missiles and Space Division.

on the DISCOVERER and MIDAS satellites;

Here they find inventive and creative freedom. Here

We invite you to join their proud company.

SUNNYVALE, CALIFORNIA • SUNNYVALE, PALO ALTO, VAN NUYS, SANTA CRUZ, SANTA MARIA, CALIFORNIA · CAPE CANAVERAL, FLORIDA · HAWAII

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Thermal Open Type Relay

Clairtron Mfg. Co., Box 171, Orange, N. J., has developed a Surgistor Thermal Relay of an inexpensive open type construction. This thermal element is adaptable to many new and existing projects where an inexpensive single-pole single-throw or a single-pole double-throw relay is needed to replace magnetic type relays. The actuating coil being a resistive type element can replace a resistor in any circuit. The contacts of the device can be set to break the circuit when a surge in current exists. The circuit will remain open as long as the trouble continues, but will return to normal automatically when fault is corrected. By using a single-pole double-throw unit a warning light or buzzer can be actuated to show failure. Contact closure is of the slow make-or-break type and is suited to applications requiring a slight delay. Contact rating is 3 amperes resistive 115 vac with a minimum of 100,000 operations. These units will operate on ac or dc voltage and power drain is 2 watts with one watt available. Voltages available are 2 to

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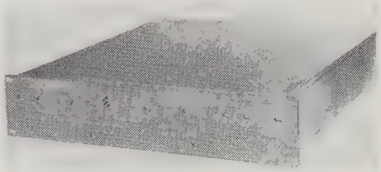
115 volts ac or dc operation. The model number is BR-101.

Price schedule for single-pole single-throw type, 1–9, \$1.25; 10–24, \$.99; 25–99, \$.86; 100–499, \$.75; 500–999, \$.70; 1000 & over, \$.65.

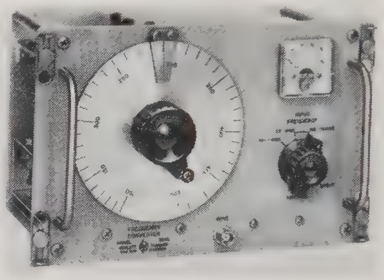
Impedance Converter



New model IC 130 impedance converter (super cathode follower), designed by **Computer Engineering Associates, Inc.,** 350 N. Halstead, Pasadena, Calif., buffers piezoelectric transducers and other high impedance signal sources from recorders, preamplifiers, amplifiers and other instruments. Bandwidth is 1 cps to 100 kc. Complete isolation of both circuitry and chassis from ground permits use of grounded or floating signal sources and recorders. Input impedance is 1000 megohms to allow low frequency response with capacitive signal sources; maintains gain accuracy when coupled to resistive signal sources. Driven inner shield of the triaxial input cable reduces effective cable capacity sensed by signal source to less than 2% of physical cable capacity. Common mode rejection of 130 db with up to 200 v from signal source and long input cables. Output capability of ±28 v at ±1 ma is ample to drive direct recorders, analog to digital converters, oscillators and other data acquisition instruments. Construction is modular type with 6 per rack, designed for 19-inch relay rack, front panel height 3½ inches. For further information write to the firm.

Frequency Converter

A new plug-in unit which increases the measuring capability of -hp- Model 524 electronic counters to 510 mc is now available from **Hewlett-Packard Co.,** 1501 Page Mill Rd., Palo Alto, Calif.



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The frequency converter unit, Model 525C, can be used in the -hp- Model 524B, C or D counters, 1) to measure frequencies between 100 and 510 mc with 100 mv sensitivity, 2) to amplify signals between 50 kc and 10.1 mc with 20 mv sensitivity. All Model 524 features are retained including stability of 5 parts in 10^8 per week for the Models 524C and D, frequency indications to 10.1 mc directly, time interval from 1 microsecond to 100 days, period from 0 cps to 100 kc and maximum resolution of 100 nanoseconds.

Model 525C contains a capacity-loaded cavity for frequency determination and a highly efficient diode harmonic generator plus a transistorized amplifier. A "go-no-go" meter on the front panel shows when the signal has enough amplitude for frequency measurements.

Model 525C is easily inserted in Model 524. It is priced at $425. Current availability is 10 weeks.

Voltage To Frequency Converter

The Magaverter, a completely static, solid state, precision analog to voltage to frequency converter has just been introduced by **Pioneer Magnetics, Inc.,** 850 Pico Blvd., Santa Monica, Calif. The Magaverter uses only rugged, solid state components (no moving parts or vacuum tubes) to produce an output square wave whose frequency is directly proportional to the input voltage.

The converter is said to maintain an input-output linearity of $\pm\frac{1}{4}\%$. They are linear to $\pm 0.1\%$ with a temperature stability of better than 50 ppm/° C.

Eight standard models provide full scale output frequencies of from 30 cps to 25,000 cps. The maximum full scale frequency is adjustable by the customer over a 4/1 range with a built-in attenuator. Standard units provide for input voltage ranges of 0–1, 0–10, 0–100 volts. Thus a model MI-1000-A will provide an output whose frequency can be set to vary linearly

(Continued on page 154A)

A STUDY FROM LIFE FOR RYAN BY D. MILLSAP

NEWS New Products

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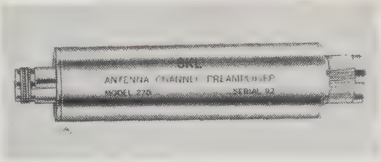
anywhere from 0 to 250 cps up to 0 to 1000 cps as the input signal is varied from 0 to 1 volt dc.

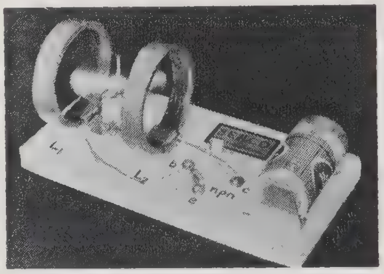
All models are assembled in a drawn steel Mil-T-27 can measuring $4\frac{5}{16} \times 3\frac{11}{16}''$ $\times 5\frac{9}{16}''$.

Channel Preamplifier



Development of a transistorized, lightweight, miniaturized television and telemetering channel preamplifier in the VHF spectrum is announced by **Spencer-Kennedy Laboratories, Inc.,** 1320 Soldiers Field Rd., Boston, Mass. Having a flat response across 6 mc, the amplifier provides 16 db of gain and has a noise figure of 9 db or less. The ampl fier is packaged in a weatherproof cylinder; over-all dimensions $1\frac{1}{4}''$ outside diameter, length including connectors $5\frac{3}{4}''$; total weight 10 ounces. Primary power is 28 volts dc supplied through the output cable. The Model 270 Channel Preamplifier is available from stock in all standard TV channels 2 through 13. Quantity inquiries for other bandwidths and at other frequencies from 220 mc to dc are invited.

Superregenerative Electronic Motor



This direct current motor, developed by Dr. Harry E. Stockman, has eliminated the brushes and commutator of the conventional dc motor, replacing them by a switching transistor. The alnico-magnet rotor spins around in either direction in a transistor-controlled magnetic field. High efficiency is made possible by operation in the superregenerative mode, with the rotor providing the self-quenching action. The photo shows a model primarily intended for educational purpose, and selling for \$14.50. It operates continuously for a month or more on a 1.5 volt flashlight cell. This motor is currently manufactured by **SER Company,** 543 Lexington St., Waltham, Mass.

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These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

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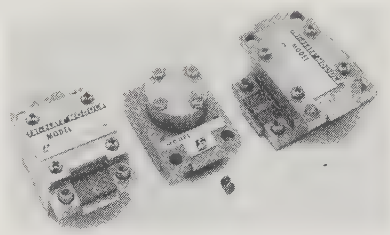
Semiconductor Mounts

Sanders Associates, Inc., 95 Canal St., Nashua, N. H., announces several new Tri-Plate® Module semiconductor mounts. Designed for use with pill tunnel diode and pill varactor packages similar to those used by Microwave Associates and Sylvania, these new mounts are complete electrical and mechanical subassemblies. The standard 50-ohm impedance model is available for immediate shipment while special impedances from 5–100 ohms are shipped within 4 weeks after the order is received.



The new semiconductor mounts for use with the $\frac{1}{8}''$ diameter, $\frac{1}{8}''$ long pill packages are the latest addition to the growing mount line. Currently, the firm offers 20 semiconductor mounts and 3 transistor mounts for off-the-shelf delivery. Such semiconductor devices as the standard crystal cartridge, modified crystal cartridge (tip removed), double-ended cartridge, miniature double-ended cartridge, miniature glass diode, pill and TO-5 and TO-18 transistor case packages can be dropped into the appropriate mount. The mount then can be used along with the rest of the Tri-Plate microwave "building blocks" for strip transmission line breadboarding.

The mounts are available in 1, 2, 3 and 4 port series and shunt connected diode models, with or without RF bypass. Series connected mounts are also available with either or both ground returns and bias injection. The shunt connected mounts can be supplied with injection terminals which are not considered ports. Two port transistor mounts are available with grounded emitter, grounded base, or 3 port mounts with emitter, base, and collector terminals connected to RF ports.

Linearly Polarized Waveguide Horn

Wiley Electronics Co., a subsidiary of Giannini Scientific Corp., 2045 W. Cheryl Dr., Phoenix, Ariz., announces a new addition to their line of microwave components. The product is a linearly polarized, high gain, high efficiency waveguide horn. Gains of 31 db and efficiencies of 70 per cent are available in this line of calibrated horns.

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These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

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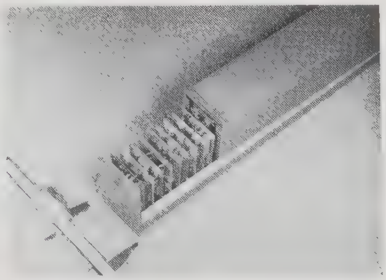
The horn design employs a square aperture with diagonal polarization. The resultant aperture illumination is tapered in all planes, which results in a circularly symmetrical pattern with extremely low side lobes and constant phase over the entire aperture.

The new calibrated horns are available with gains of 19, 25, or 31 db, and in K_a, K_u, or X band.

Wiley also manufactures high speed waveguide switches, waveguide sector scan twist sections, and complete microwave antenna systems, in addition to a new line of subminiature telemetering equipment.

Digital To Analog Converters

A new line of solid state digital-to-analog converters designed by **Dynamic System Electronics,** 2001 N. Scottsdale Road, Scottsdale, Ariz., for military and industrial applications in hybrid digital-analog systems is being offered with a 300 KPPS maximum clock rate. The unit is self-contained including storage, power supply, reference transformer, and forced air cooling. Power requirements are 40 watts, 60 to 400 cps, 115 volts; and the reference voltage is 115 volts, 60 cps or 400 cps, 1 watt. Six units can be mounted in a standard 19 inch rack requiring $3\frac{3}{4}$ inches of panel height. Delivery in quantity can be obtained within 60 days.



Either ac or dc models up to and including ten bits and a sign bit are also available. The modular design of these units allows economical modifications to particular requirements. For details contact the firm.

(Continued on page 160A)

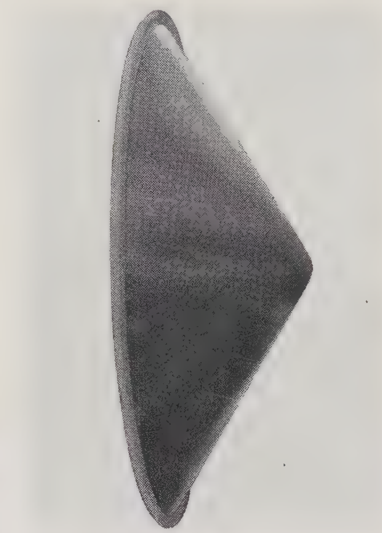
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 158A)

Parabolic Antenna Covers

Tower Construction Co., 2700 Hawkeye Dr., Sioux City, Iowa, has devised an inexpensive means of protecting parabolic antennas from severe winter weather by using an antenna cover, which is cone shaped and made of molded fiberglas.



According to Tower engineers, the fiberglas Para-Dome costs from 35% to 50% less than other protective type covers.

The Para-Dome has been designed to withstand wind force of 50 lbs. per square foot. Tests have proved that signal attenuation between 2000 and 6200 mc is within 0.1 db.

The Para-Dome is available for use with reflectors of 4-, 6-, 8-, and 10-foot sizes. Special sizes are available on request.

Normal delivery on standard sizes is fifteen days.

Additional information can be obtained by writing to the firm.

Transducer Bulletin

Capabilities of a new Type 4-329 airborne pressure transducer are described in a two page bulletin available from **Transducer Div., Consolidated Electrodynamics Corp.**, a subsidiary of Bell & Howell Co.

The 4-329 is a medium-to-high range unit measuring pressures from 0-100 psi to 0-5000 psi absolute. Operable temperature range is $-100°$F to $+300°$F.

Bulletin 4329 may be obtained from CEC, 360 Sierra Madre Villa, Pasadena, Calif.

Gallium Arsenide

Gallium Arsenide in single and polycrystal forms for use in tunnel diodes, varactor diodes, microwave diodes and transistors is now available from **Alloys Unlimited Chemicals, Inc.**, 42-73 Hunter St., Long Island City 1, N. Y.

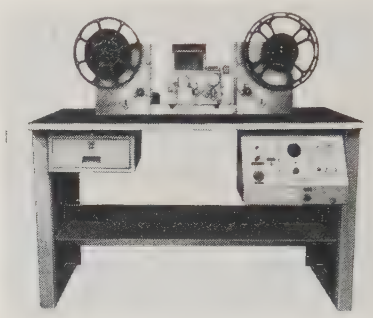
The compound is available in resistivity ranges of 0.0X to 0.000X Ohm-Cm for p-type devices and 0.X to 0.000X Ohm-Cm for n-type devices. It has a carrier concentration of 4×10^{16} carriers/cm³ to 5×10^{19} carriers/cm³ and mobility up to and better than 4500 cm² volt/second.

The compound can be supplied undoped or doped with either zinc, cadmium, manganese, tellurium or tin.

Alloys Unlimited produces semiconductor compounds in polycrystalline ingots, single crystals and microcrystalline powders.

Tape Editing Table

A new high-speed editing table and viewer for rapid film scanning has been introduced recently by **Camera Equipment Co., Inc.**, 315 W. 43rd St., New York, N. Y.



Called the "CECO high-speed editing table," this new unit is available in both 16mm and 35mm models. It is suitable for television stations and film libraries where rapid inspection of prints, insert spots, and commercials 250 feet per minute in both forward and reverse is required before release. It has the ability to stop the film action on a single frame without damage to the film.

Built to government specifications, the viewer has a 4″×6″ screen and comes equipped with a footage counter and optical sound head.

A panel contains speaker, amplifier and controls for both speed and sound. All components are mounted on a formica top table with steel frame, film storage rack and drawer.

Accessory equipment includes fluorescent illuminated film clip racks, utility drawers, plastic covered light wells and racks for short rolled film clips. A time counter is also available at additional cost.

Maximum dimension: 60″ long, 30″ deep. Table height, 36″. Total height with 3000 foot reel: 54″. Price of the 16mm table is $2,500.00. The 35mm table is $1,750.00. Accessories are additional.

(Continued on page 162A)

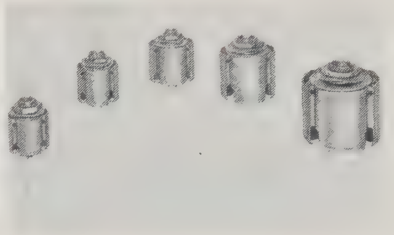
NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 160A)

Adjustable Cores

Ferroxcube Corp. of America, Saugerties, N. Y., announces a new line of Ferroxkor adjustable pot cores for filtering applications with frequencies up to 3 mc.



These cores are manufactured to close tolerances of effective permeability, making it possible to precalculate the inductance of the wound coil to within ±3%. Each pot core assembly has an adjustment range of ±7% of the mean value of permeability allowing adequate compensation for wide tolerance capacitors. Final adjustment and readjustment, should disassembly be required, can be easily effected to an accuracy of better than 0.02%. This range and accuracy is accomplished by means of the simple, self-locking adjuster that is built into the core. Unique design and construction facilitate the rapid assembly of inductors under conditions of large scale production. Available in 5 different sizes with a choice of 3 standard effective permeabilities. Complete details can be obtained by writing the manufacturer.

Rotary Indicators

Daco Instrument Co., Tillary and Prince Sts., Brooklyn 1, N. Y., has designed and is in production on a line of micro-miniature rotary indicators. These indicators come in two sizes, the smaller of which measures 0.375″ diameter by 0.562″ long. Designed to use 100 milliwatts or less where required, the component weighs 3.7 grams. Supplied for any standard voltage to 30v dc or ac at 400 cps. Also available for 110v operation with an external resistor. Rotation can be up to 60° either C.W. or C.C.W. The solenoids operate in a temperature range of −65° to 165°F, are of rugged construction with jewel bearings and have a standard extended shaft of 1/32″ diameter by 3/16″ long. Other shaft lengths are available. These electrical indicators which signal by means of a shutter arrangement, can be used as a malfunction indicator, annunciator, binary readout in computers, output indicators in transistor circuits and in any other application where a two-position indication is required.

(Continued on page 164A)

THE PRACTICALITY OF PHOTO-ETCHED MICROWAVE CIRCUITS

with particular emphasis on loss and power handling characteristics.

by JAMES W. CHRISTIAN
Project Engineer, Instrument Division
LABORATORY FOR ELECTRONICS, INC.

Although the advantages of printed circuit techniques in the lower frequency regions are well known, it is not so well known that printed circuits at microwave frequencies now enjoy the same advantages: light weight, small size, reproducibility, and economy. Even further, these multiple advantages of photo-etched transmission line now permit routine solutions to transmission line and antenna problems once considered virtually impossible by conventional waveguide or coaxial line techniques.

In applying photo-etched techniques to microwave circuits, we occasionally find that prospective users believe that the losses may be too high and power handling capabilities too low. Such beliefs are not true for most applications. We hope that the graphs shown, particularly those of attenuation and power handling capacity, will dispel these impressions.

The planar transmission system used in photo-etched applications evolved from the coaxial line. As shown in figure 1, if a coaxial line is deformed until the outer and inner conductors become long rectangles, then removal of the side walls results in a flat transmission line. Its form factor is ideally suited to printed circuit techniques.

Figure 1

This planar line is substantially free from radiation losses. Almost all of the field is concentrated in the region of the inner conductor, and since the two outer conductors are at the same potential, the plane of the center conductor is an essentially field-free region. As in a coaxial line, the TEM mode is used in practical applications.

LFE's photo-etched components are called Epsi-line, a contraction of Epsilon (from dielectric constant) and transmission line. However, Epsi-line is more than a trade name. It includes true production engineering and outstanding features such as an outer encapsulation that provides > 120 db attenuation to radiation pickup from outside sources.

Physically, Epsi-line is made from thin copper-clad laminates, usually 1/16″ thick, wherein unwanted copper is removed by photo etching. The space between inner and outer plates is filled with a low-loss dielectric material, and hence components become much smaller from this fact alone. The wave length in the line, λg, is shortened almost 40% since $\lambda g = \frac{\lambda o}{\sqrt{\epsilon}}$, where λo is the free space wave length and ϵ is the dielectric constant. ($\epsilon = 2.53$ for the commonly used laminate.) Also, circuit elements may be placed close together without interaction. Cross talk at 4000 mc/s is roughly 70 db per 1/4 inch.

CHARACTERISTIC IMPEDANCE

The characteristic impedance of Epsi-line is determined by the dimensions of the inner conductor, the dielectric constant of the laminate, and the spacing between the ground planes. Since these can be controlled very accurately, homogeneity of impedance is assured, and internal junction VSWR's of < 1.05 are readily maintained. Note in figure 2 that a wide range of impedance levels is practicable.

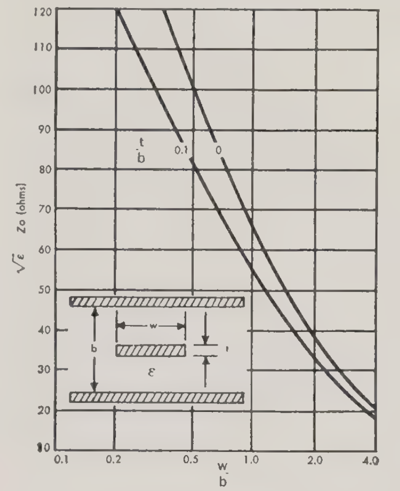


Figure 2

ATTENUATION

As in all transmission lines, the attenuation or loss factor is composed of three parts: losses in the conductors, in the dielectric medium, and by radiation. But since losses by radiation and losses in the ground planes are extremely small (10^{-3}) compared with the other losses, attenuation can be expressed as the sum of inner-conductor and dielectric losses. For average dimensions of center conductor and for Rexolite 1422 dielectric, the attenuation per foot of line is given in figure 3. Only a very few inches of line are used in any component. Consequently, even at X-band, the insertion loss of Epsi-line components is less than 0.4 db.

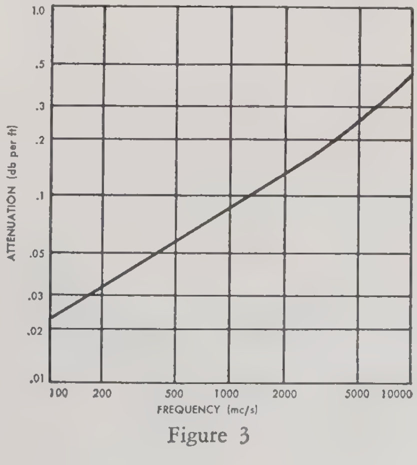


Figure 3

POWER RATING

As one might expect, average power rating depends on the allowable temperature rise of the inner conductor. It is also interesting that the temperature rise in the inner strip of Epsi-line is nearly the same as in an ordinary coaxial line having the same characteristic impedance and attenuation per unit length.

For a temperature rise of 39°C, $Z_0 = 50$ ohms, $t = 0.002''$, $w = 0.093''$, $b = 0.125''$, $\epsilon = 2.53$ (Rexolite 1422), figure 4 gives the average power capabilities of representative Epsi-line components. At X-, C-, S-, and L-bands, average power capabilities are approximately 100, 150, 200, and 500 watts, respectively.

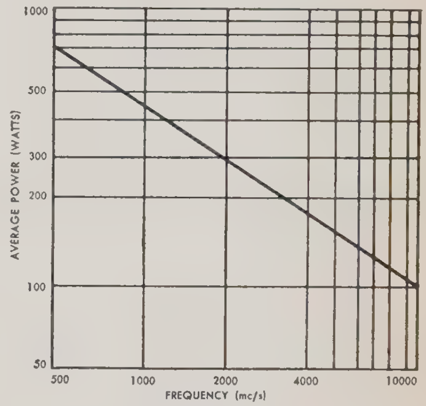


Figure 4

The peak-power capability of standard configuration Epsi-line is restricted by a small air gap at the edge of the photo-etched center conductor. Under the same conditions and dielectric mentioned above for average-power rating, the peak power rating is 25 kilowatts with a 2:1 safety factor. However, the air gap can be removed by special processing techniques. This plus appropriate internal encapsulation permits peak ratings as high as 1.5 megawatts.

With ratings such as these, Epsi-line microwave components are truly practical. In the frequency range 500 to 11,000 mc/s, LFE manufactures a wide variety of cataloged components: balanced crystal mixers, directional couplers, hybrids, attenuators, multiterminal power dividers, as well as any combination of such devices in a single package. Input VSWR's are generally less than 1.2 over wide frequency ranges. The total encapsulation, internal and external, assures constant peak performance protected from radiation, moisture, corrosion, shock, and vibration. For further information, write to Mr. Perry Pollins at the address below.

LABORATORY FOR ELECTRONICS, INC.
Instrument Division
714 Beacon Street, Boston, Massachusetts

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 162A)

Demagnetizer

This demagnetizer, a product of **Amplifier Corp. of America,** 398 Broadway, New York 13, N. Y., produces complete erasure of recorded signal on all brands of tape and $\frac{1}{4}''$, $\frac{1}{2}''$, or 16 and 35 mm. magnetic sound film—on plastic or metal reels of any size from 5″ to 15″.

The erasing process is so efficient that, even on severely overloaded tape, the background noise level is lowered 3 to 6 db below that of unused tape. Since erasure is effected on the reel—without unwinding and rewinding—there is no wear on the tape against erase heads. The device is portable, and applicable to demagnetizing record-playback and erase heads, reducing background noises and tape distortion.

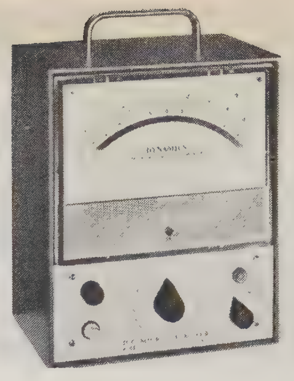
This demagnetizer operates on any alternating (50 or 60 cps) current, and furnishes the necessary gradually diminishing cyclic magnetization field which the tape normally encounters during supersonic erasure. To erase you place the demagnetizer on top of, and move it around the reel of tape. Within seconds, the completely erased reel is ready for re-use. Application to new, unused tape will help produce a recording with a greater signal-to-noise ratio.

Called the "Magneraser," this demagnetizer is available in two models: Model 200C for 100–130 volts; Model 220C for 200–260 volts.

Additional information may be obtained by writing to the manufacturer.

DC Microvoltmeter

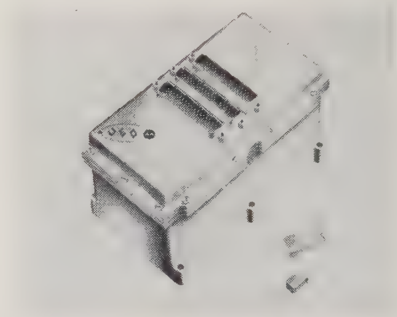
A dc microvoltmeter from **Dynamics Instrumentation Co.,** 583 Monterey Pass Rd., Monterey Park, Calif., offering complete freedom from power-line coupling by means of an automatically-recharged nickel-cadmium battery, the Model 4472 has 15 ranges from ±100 microvolts full-scale to ±1000 volts full-scale. The input impedance is 100 megohms or greater on all ranges except that it is 10 megohms or greater on the microvolt ranges.



The all-transistor instrument has ±1% accuracy on all ranges, a 7.2″ mirror scale, and substantial overload protection on all ranges. Model 4472 is a portable instrument and Model 4472R is the rack-mounted version.

Electronic Commutator

A low-level electronic commutator has been developed by **United ElectroDynamics, Inc.,** 200 Allendale Rd., Pasadena, Calif.



Contact resistance averages 12 ohms. Reverse current is less than one nanoampere per channel, and special shielding has achieved a common mode rejection of better than one million to one.

The commutator will operate reliably and within specification at +125°C.

Solid State Digital Converter

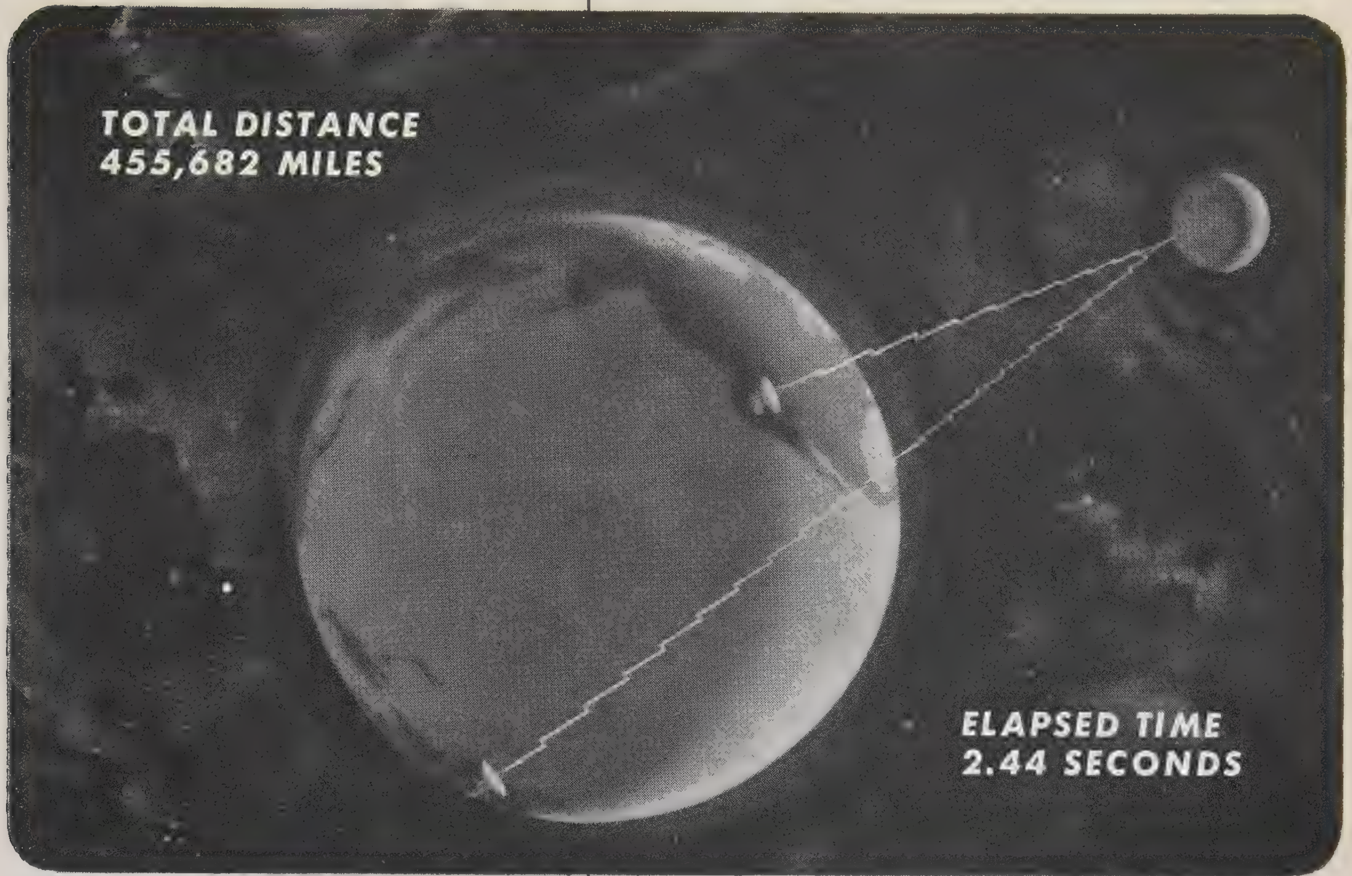
Winsco Instruments & Controls Co., 11789 W. Pico Blvd., Los Angeles 64, Calif., manufacturers and designers of precision instruments and controls, have introduced an all solid-state digital converter, Model 4310 which provides a simple accurate means for direct digital measurement of temperature, pressure, and other process variables.

Used with digital frequency-period counters, the Model 4310, converts the output of potentiometric transducers into a variable frequency which can be displayed or recorded digitally on the counter or counter-printer. Small size and low

(Continued on page 166A)

Notable Achievements at JPL

MOON BOUNCE... a collaborative project of the National Aeronautics and Space Administration, the Jet Propulsion Laboratory, and the Australian Ministry of Supply to link two continents by radio signals bounced off the Moon



On February 10, 1961, California and Australia were linked in the first international space communication experiment that bounced voice messages between the two points via the Moon. The words were beamed at the Moon from the Jet Propulsion Laboratory transmitter at Goldstone, California to the receiver at Woomera, Australia.

Principals in the conversation were Dr. Hugh L. Dryden, NASA Deputy Director, whose voice was relayed from Washington by telephone; Dr. Lee DuBridge, President of California Institute of Technology, who spoke directly from Goldstone; and Alan Hulme, Australian Minister of Supply at Woomera.

The occasion tested the new Australian station, the second of three Deep Space Instrumentation stations developed and directed for the National Aeronautics and Space Administration by the Jet Propulsion Laboratory.

CAREER OPPORTUNITIES AT JPL
IN THESE FIELDS—NOW

Electronic Engineers

. . . for component and system design of deep space communications, instrumentation, and automatic control equipments.

. . . for microwave and RF solid state circuit design and flight evaluation.

. . . for project management assignment on advanced development and contracted effort in space communications.

Physicists

. . . for analysis in communications theory, orbital mechanics, guidance and control, and systems performance.

. . . for analysis of digital communication and control systems; real-time digital computer and closed-loop systems.

. . . for research and development of servo and control mechanisms for large ground based and spacecraft antenna systems.

•

Other opportunities exist for electronic engineers and physicists in many areas at JPL which has been assigned the responsibility for the nation's Lunar, Planetary and Interplanetary unmanned exploration programs.

CALIFORNIA INSTITUTE OF TECHNOLOGY
JET PROPULSION LABORATORY
PASADENA, CALIFORNIA

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 164A)

power consumption are inherent in the solid state design of the unit.

Operating directly into a transmitter, the converter is an inexpensive, low power requirement, telemetry signal generation system. For multi-channel operation, it can be connected directly to the input of the transmitter for transmission of several data channels over a single data link.

Operating characteristics: Input from any resistance source within 500 to 3000 ohms; full range output frequency deviation, 7.5% at 5 vac; models available for the frequency range 400 cps to 70 kc and for all standard telemetry channels; conversion linearity, +1% for 7.5% frequency deviation; low temperature sensitivity of 15 ppm/°F; power requirements, 0.95 ma, 21 vdc ±15%; with a sensitivity of 80 ppm frequency change per 5% change in supply voltage. Lightweight (4 oz.), and small (1.5″ diameter×2.0″ high), the converter is priced at $159 in single lots.

For additional information, contact Emo D. Porro at the firm.

Neal Appointed By ACF

The appointment of Harold T. Neal as Paramus (N. J.) plant manager for the electronics division of **ACF Industries, Inc.,** effective immediately, is announced by George B. Shaw, division general manager.

Neal comes to ACF from Cook Electric Co. of Chicago, where he was vice president in charge of the Trans-Digital Systems division. Before joining Cook in 1958 he was with Stromberg-Carlson, a division of General Dynamics, as business manager for electronics systems.

Neal is a lieutenant colonel in the Air Force Reserve, having twice served on active duty, once during World War II and again from 1947 to 1956.

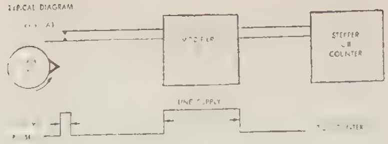
He holds a bachelor of science degree in chemical engineering, from Syracuse University, and a master of business administration degree in management of research and development, from the University of Chicago.

Pulse Modifier

A new pulse modifier designed and developed by the **Programation Division, Guardian Electric Mfg. Co.,** 1621 W. Walnut St., Chicago, 12, Ill., provides a means for driving standard electro-mechanical devices, such as counting units, from a high speed, light duty, low current, pilot contact.

PULSE MODIFIER

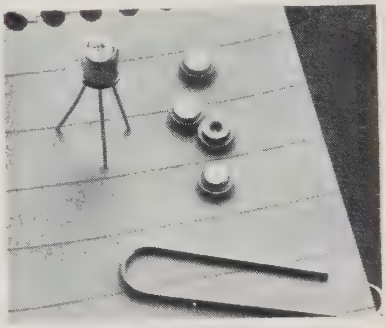
FIXED OR VARIABLE TYPES



The pulse modifier is of modular construction, solid state, plug-in type. An example of its usage would be a stepper or rotating cam making and breaking a set of contacts within 1 millisecond, taking this 1 millisecond pulse of low energy (50 milwatts) and modifying it to operate a stepper or counting unit requiring 15 milliseconds and 15 watts of power to operate. Pulse modifiers are available in fixed or variable models. Variable types allow for variations in pulse duration and wattage requirements. Line supply as specified. Outputs to 15 watts and 10–100 ms, 24 vdc. Special units can be made to order upon receipt of specific requirements as to wattage and output millisecond pulse width. For further information write Jerome S. Winkler, Programation Div.

Tunnel Diodes

The General Electric Co., Liverpool, N. Y., has announced four new microwave frequency germanium tunnel diodes housed in a new miniature package.



Featuring tightly controlled, low peak currents, their resulting high negative resistance better answers the impedance matching requirements of many microwave systems.

The new microwave devices are also designed for application in radar, very high frequency amplifiers and oscillators and other "S" band equipment.

They are immediately available in production quantities and are priced from $20.00 to $30.00 each to original equipment manufacturers.

Two of the new germanium tunnel diodes operate at frequencies up to 3500 mc and the other two at frequencies up to 4600 mc.

These components are suitable where extremely low inductance is required. The typical inductance of the new devices is 0.4 millimicrohenries. This parameter is held to a maximum of 0.5 millimicrohenries. By comparison, this inductance is about one-tenth that normally found in germanium tunnel diodes housed in the TO-18 case.

(Continued on page 168A)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 166A)

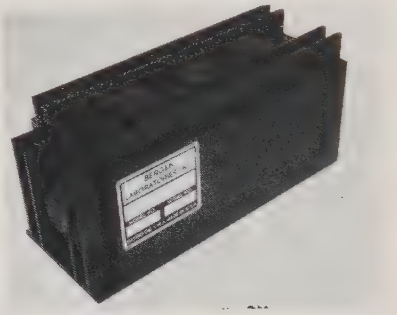
Housed in a hermetically sealed "strip-line" package, the new high speed devices have been used to replace traveling wave amplifiers, masers, and parametric amplifiers.

They have received JEDEC designations 1N3218, 1N3218A, 1N3219 and 1N3219A. The devices have typical total capacities of 7, 4, 14 and 7 picofarads, respectively.

Typical peak point current ratings of the 1N3218 and 1N3218A are 1.0 milliamperes and of the 1N3219 and 1N3219A are 2.2 milliamperes. This parameter rating is controlled to within ±10%.

Power Supplies

A line of inverters, converters and power supplies has been developed by **Bergen Laboratories, Inc.,** 60 Spruce St., Paterson, N. J., which features small size, high temperature operations and reliability.

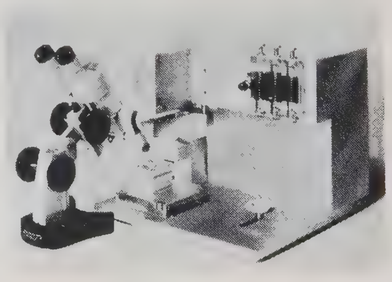


Particularly featured are three phase sinusoidal inverters for missile and aircraft use, ranging in power from 15 va to 300 va three phase (delta or wye). Distortion is held to below 5% and regulation for line and load from 1% to 5%.

Some of the units make use of SCR type inversion, believed to be the highest efficiency means of conversion.

Semiconductor Lead Bonding Machine

A new thermal compression lead bonding machine for semiconductor manufacturing is announced by **Diotran Pacific,** 101. Alma St., Palo Alto, Calif.



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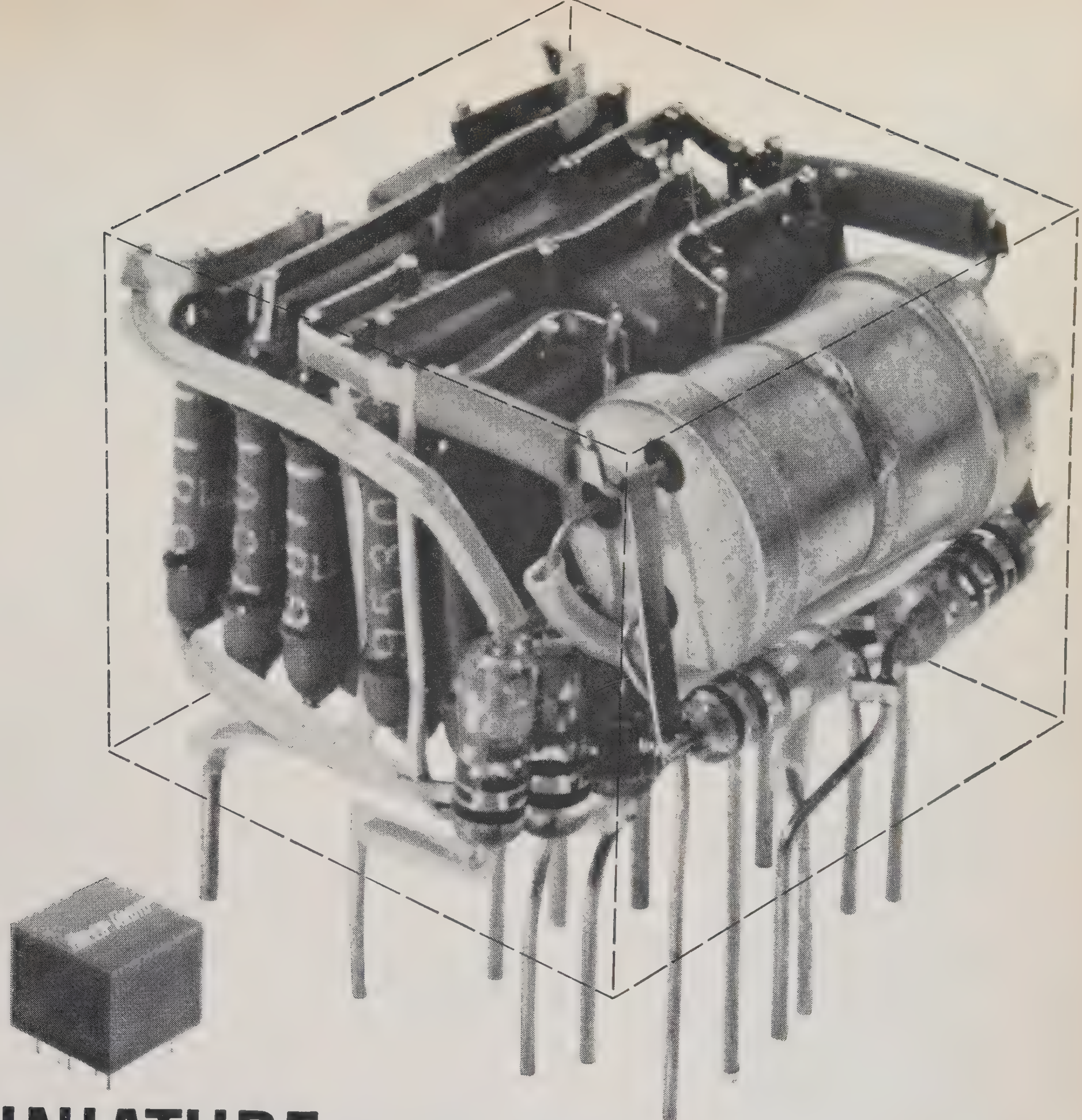


MINIATURE BUILDING BLOCK MODULES

5 times actual size to better show the 28 standard-sized components

Delco Radio's high density packaging of reliable standard components utilizes the unique three-dimensional welded wiring technique. These miniature modules are available off the shelf in 16 basic types. Or with them, Delco Radio can quickly build for you a compact, reliable digital computer for airborne guidance and control or any other military application. Vacuum encapsulated with epoxy resin, the modules perform all the standard logic functions. They meet or exceed all MIL-E-5272D (ASG) environmental requirements, and operate over a temperature range of −55°C to +71°C. Too, these same reliable digital circuits are available packaged on plug-in circuit cards. And we can also supply circuits to meet your specific needs. For complete details, just write our Sales Department. *Physicists and electronics engineers: Join Delco Radio's search for new and better products through Solid State Physics.*

PIONEERING ELECTRONIC PRODUCTS THROUGH SOLID STATE PHYSICS

DELCO RADIO

Division of General Motors • Kokomo, Indiana

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 168A)

The lead bonder provides fast, accurate bonding of gold wire to semiconductor materials. High precision micro-manipulation, excellent optical system, and high output heater assembly are said to insure one-time, positive bonding and substantially reduce rejects.

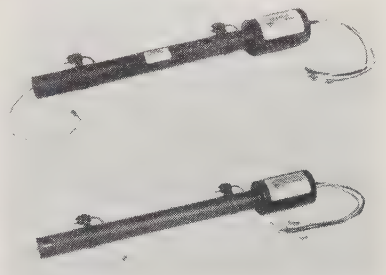
Gold wire and wire housing are protected by gas-filled envelope. Built-in, precision flame cutting tip severs and "balls" wire cleanly.

The unit requires minimum operator training time, keeps efficiency high. Simplicity of design makes possible low initial cost and low in-service maintenance. Unit is compact. Total bench space requirement, including optical system, is 13 inches by 15 inches.

Complete details are available from the manufacturer.

Traveling-Wave Tubes

The development of two low-cost permanent-magnet focused traveling-wave tubes has been announced by **Sylvania Electric Products Inc.,** 1100 Main St., Buffalo 9, N. Y., a subsidiary of General Telephone & Electronics Corp.



Sylvania's Microwave Device Operations say the new L-band amplifiers are part of a series which will cover the 1,000- to 12,000-mc frequency range. The tubes employ integral permanent magnets to reduce their size and weight without sacrificing electrical performance. Type TW-4267 ($3\frac{1}{2}$ pounds) provides over 15 milliwatts RF output power and 35 db gain from 1,000 to 2,000 mc. Type TW-4268 (4 pounds) operates over the same range with more than 1 watt RF output power and 30 db gain. The tubes have a maximum diameter of $2\frac{1}{4}$ inches.

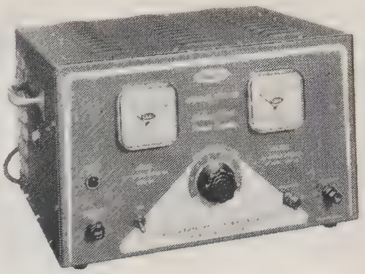
The low cost and compact design of the new components suit them for test equipment applications. In addition, their reduced size and weight "make them superior to bulky solenoid types in the design of compact, wideband equipment."

Small quantity price for both types is $925. Evaluation units available immediately, and small quantities can be delivered in approximately 60 days.

Additional information on TW-4267 and TW-4268 may be obtained from local Sylvania sales offices or from Buffalo.

DC Power Supply

A new low ripple filtered 12-volt dc power supply, designed for servicing mobile electronic equipment and serving as a shore dc power source for boats, has been developed by **Electro Products Laboratories,** 4500 N. Ravenswood Ave., Chicago, 40, Ill.



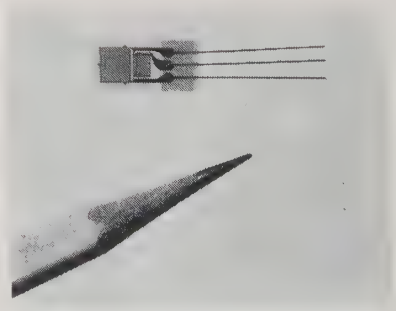
The new PS-30 power supply is nominally rated at 12 volts and features low ripple (less than 1% at 30 amperes). Operating on 115 volts, 50/60 cps input this unit provides an output current of 30 amperes when operated continuously and up to 50 amperes if operated intermittently for short periods. The PS-30 also features low output impedance.

Patented conduction cooling provides a greater margin of safety to both equipment being tested and the unit. At the same time it increases the rectifier life and lowers the cost per ampere output. Special circuitry and use of top quality components are claimed to increase the performance life and enable it to withstand heavy overloads for long sustained periods of operation.

For complete specifications on the Model PS-30 write to C. J. West, Electro Products Laboratories.

Strain Gage With Computer

A new bonded resistance-foil strain gage, with a built-in computer that solves general strain-to-stress equations automatically, has been developed by **Electronics & Instrumentation Div., Baldwin-Lima-Hamilton Corp.,** 42 Fourth Ave., Waltham 54, Mass.



The new gage will simplify the task of obtaining stress readings in a wide range of testing and measuring applications by automatically eliminating the need for calculations of stresses from the strain indications.

The SR-4 provides electrical responses

(Continued on page 172A)

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 170A)

which are proportional to either stress or strain, at the discretion of the user, by using two independent axial strain-sensing elements oriented 90° apart.

One element measures the conventional strain. The other element acts as the automatic computer by rejecting the axial component of strain caused by stress in a transverse direction. The combined elements then respond only to that component of strain which is produced by stress in the axial direction.

The two sensing elements have a common electrical connection to permit independent use for measuring conventional axial or transverse strains, or combined use to measure the stress along the principal gage axis.

Transistor Literature

The **Lansdale Div., Philco Corp.,** Church Rd., Lansdale, Penna., has published a "Transistor Guide for Communications Circuit Designers."

The guide presents a comprehensive summary of the basic ground rules to be followed in designing transistorized communications circuits. Valuable rules of thumb are included to provide circuit designers with a practical understanding of the results that can be expected from transistorized communications equipment.

Introductory paragraphs are devoted to an elucidation of the various parameters which specify and categorize the characteristics of communications transistors.

Practical information, circuit diagrams, curves and equations are given to aid in the design of transistor circuitry for audio, video, RF and IF amplifiers, mixers and converters, oscillators, multipliers, and high frequency-high power output stages.

The basic aspects of circuit configurations, variation of noise figure with frequency, biasing, and dc stability are separately examined.

Specific attention is given to short circuit impedance measurements, multi-purpose circuit functions, AGC characteristics, temperature control, and overload and cross modulation distortion.

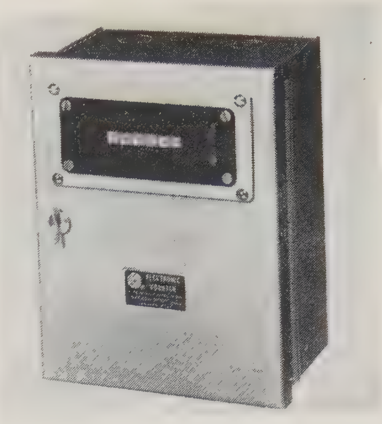
A chart is included listing specific transistors which have been designed and specified for each of the communications sockets.

The complete guide will be mailed free of charge upon request made under company letterhead.

Counting and Control

Two simple, low-cost, general purpose photoelectric counting and control systems were introduced at the IRE show by the **Electronic Controls Div., Veeder-Root Inc.,** Danvers, Mass.

Featuring a power-type photocell that directly activates a sensitive relay, eliminating the need for amplification, this new equipment enables the rapid tabulation of boxes, cartons, units, merchandise, people, etc., at areas remote from the readout device, or the control thereof where counting is not required.



The first system is a photoelectric counter ("Series 1810," see photograph), which can register up to 700 counts per minute. Minimum "Light-Off" and "Light-On" times are 30 ms. The 115 volt ac counter within the unit is operated directly by the sensitive relay.

Integral to the system is the Photohead-exciter ("Model PL"). The light source, a lamp of 6 to 8 volt, 15 candle power, is mounted in a socket behind a single lens cemented over an inlet. The power-type photoelectric cell is placed near the rear of another housing. Each is built into standard electrical OLB conduit fittings, with eight-inch leads.

The second system is a photoelectric control ("Series 1811") to provide functional actuation of auxiliary equipment through a set of internally-mounted relays. It may be adjusted to extend the hold-in time of the relay from 40 ms to one second after the light beam is restored, enabling the control to act as a jam detector in applications requiring one second or less timing to provide for the elimination of two pulses from an object with protruding parts. It is also with the Model PL Photohead-Exciter.

Both the counter and the control operate on a line voltage of 105–125 volts, 60 cps, under a maximum ambient temperature of 130°F. They each measure $8\frac{3}{8}''$ wide $\times 8\frac{5}{16}''$ high $\times 4\frac{1}{4}''$ deep, are finished in commercial gray, metalized, and have interior terminal strips for connecting.

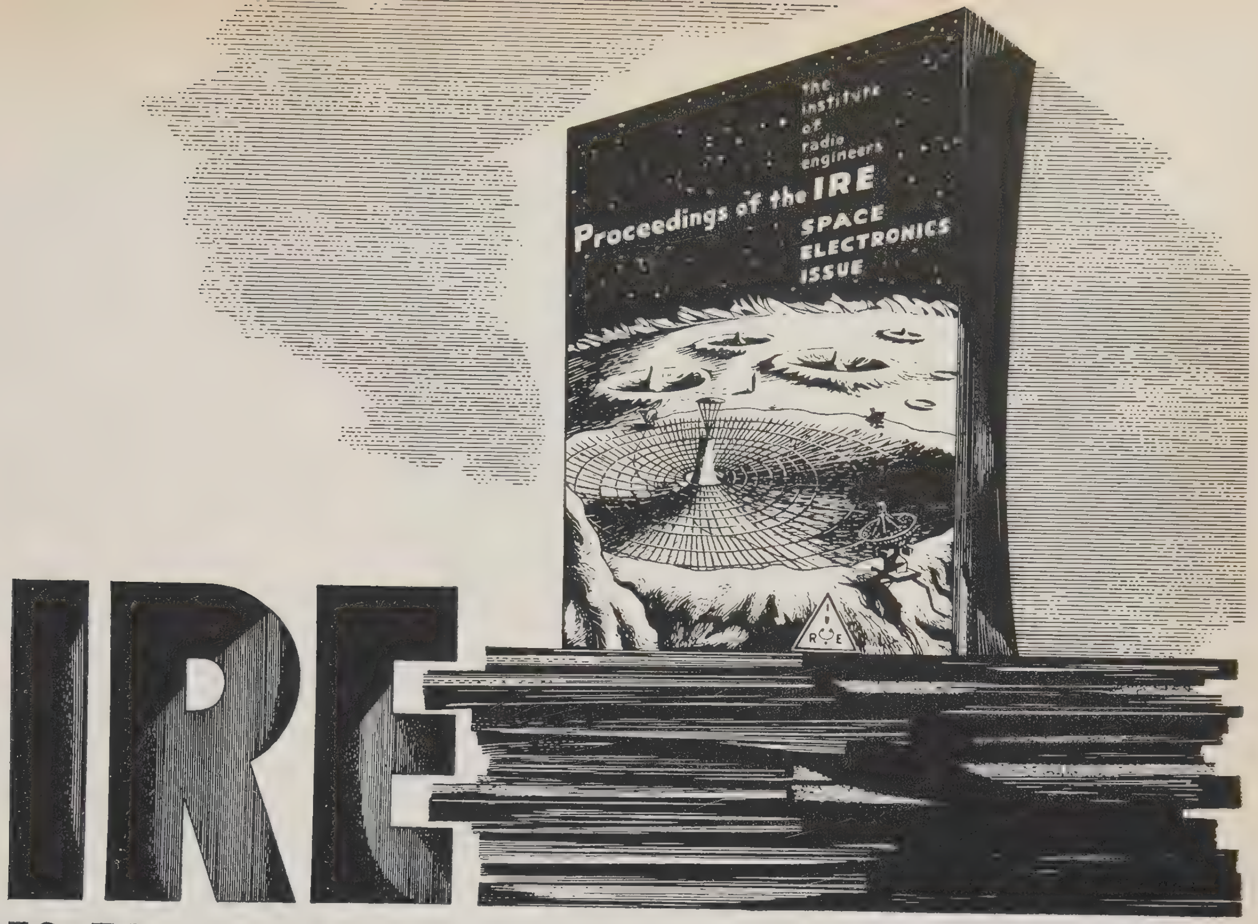
For additional information contact firm.

Gallium Arsenide Mesa Varactor Diodes

Types MS262 through MS266 are now included among the microwave semiconductor products available from **Micro State Electronics Corp.,** 152 Floral Ave., Murray Hill, N. J.

These diodes, with cutoff frequencies extending to 150 kmc, are designed for use in parametric amplifiers, microwave switches and harmonic generators. Gallium arsenide varactor diodes with even higher cutoff frequencies will also be available

(Continued on page 174A)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PROFESSIONAL CONSULTING SERVICES

NEWS New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 172A)

from Micro State within a short period of time.

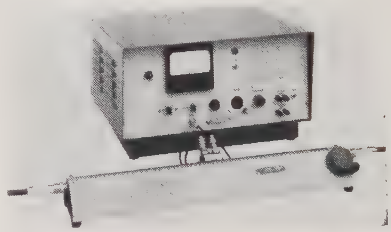
Packaged in a "pillbox" configuration 0.120″ in diameter and 0.050″ high, these gallium arsenide diodes have a typical parasitic inductance of 0.3 millimicrohenrys and maximum total capacitance of 1.5 $\mu\mu$f. Total capacitance includes case capacitance and diode capacitance. All units are rated at 6V minimum breakdown at 10 ma.

The excellent performance of these varactor diodes verifies the superiority of gallium arsenide material for use in high dissipation varactor diodes.

For additional information, write to the firm.

Direct Reading Microwave Phase Meter

Wiltron Co., 717 Loma Verde Ave., Palo Alto, Calif., announces a direct reading microwave phase meter for checking the relative phase between two signals in the 300 mc to 4000 mc frequency range. This laboratory instrument measures phase on a meter with 0.1° resolution at the microwave frequency. The principle of operation is that of square law detector response in a standing wave pattern, the standing wave pattern being the resultant of the combination of the two signals whose relative phase is being measured.



The accuracy and sensitivity of measurement which permits phase angles to be resolved to 0.1° is an important factor and compares quite favorably with even audio frequency phase meters.

The better than 10 to 1 frequency coverage and the ability to handle signals with unequal amplitudes means that this one instrument will be quite general purpose in its field and should find its place in most microwave laboratories.

The meter offers a "Servo Output" for automatic feed-back phase control. This means that in addition to serving as a metering device it can serve as an element of a phase correcting system.

The phase meter can be adapted for automatic swept frequency phase measurement with recorder output. This should be an important usage in production test applications.

The Model 300 is now available with 6 week delivery and a price of $2500. Literature is available on request.

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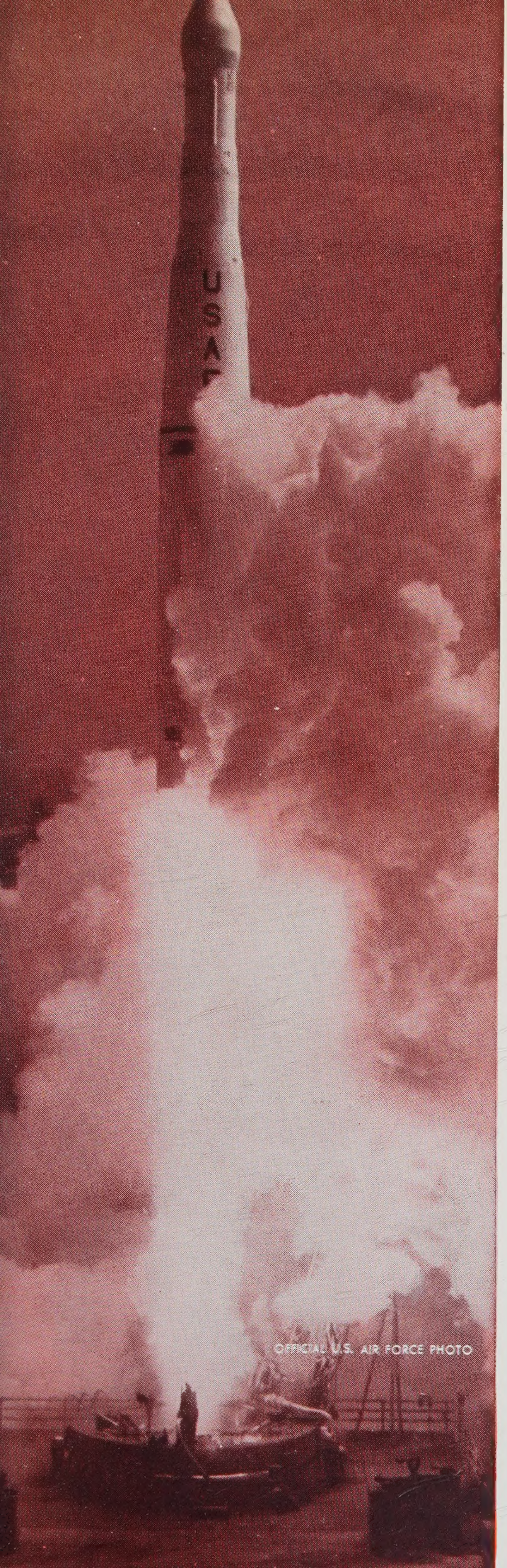

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Look to the pioneer producer of ***many millions of Military transistors*** for your critical applications. Philco has been the symbol of reliability from one of the industry's first JAN types (2N128) to the present broad Military line.

The enviable record of ultra-reliable performance has resulted in the use of Philco transistors in many Military programs. The following types are available to existing Military specifications:

TYPE NO.	APPLICATION	MILITARY SPEC. NO.
2N128	High frequency amplifier	MIL-T-19500/9A
2N240	High speed switch	MIL-S-19500/25A
2N393	High speed switch	MIL-S-19500/77A (Sig C)
2N495	Medium frequency amplifier (Silicon)	MIL-T-19500/54A (Sig C)
2N496	Medium speed switch (Silicon)	MIL-S-19500/85 (Sig C)
2N499	VHF amplifier	MIL-S-19500/72A (Sig C)
2N501A	Very high speed switch	MIL-T-19500/62 (Sig C)
2N502A	VHF amplifier	MIL-S-19500/112 (Sig C)
2N599	Medium speed switch	MIL-S-19500/166 (Navy)
2N1118	Medium frequency amplifier (Silicon)	MIL-S-19500/138 (Sig C)
2N1119	Medium speed switch (Silicon)	MIL-S-19500/139 (Sig C)
2N1158A	VHF oscillator	MIL-S-19500/113 (Sig C)
2N1199A	High speed switch	MIL-S-19500/131 (Sig C)
2N1200	HF amplifier (Silicon)	MIL-S-19500/105 (Sig C)
2N1201	HF amplifier (Silicon)	MIL-S-19500/101 (Sig C)
2N1411	High speed switch	MIL-S-19500/133 (Sig C)
2N1499A	High speed switch	MIL-S-19500/170 (Sig C)
2N1500	Very high speed switch	MIL-S-19500/125 (Sig C)

For information on any of the above types, write Dept. IRE561.

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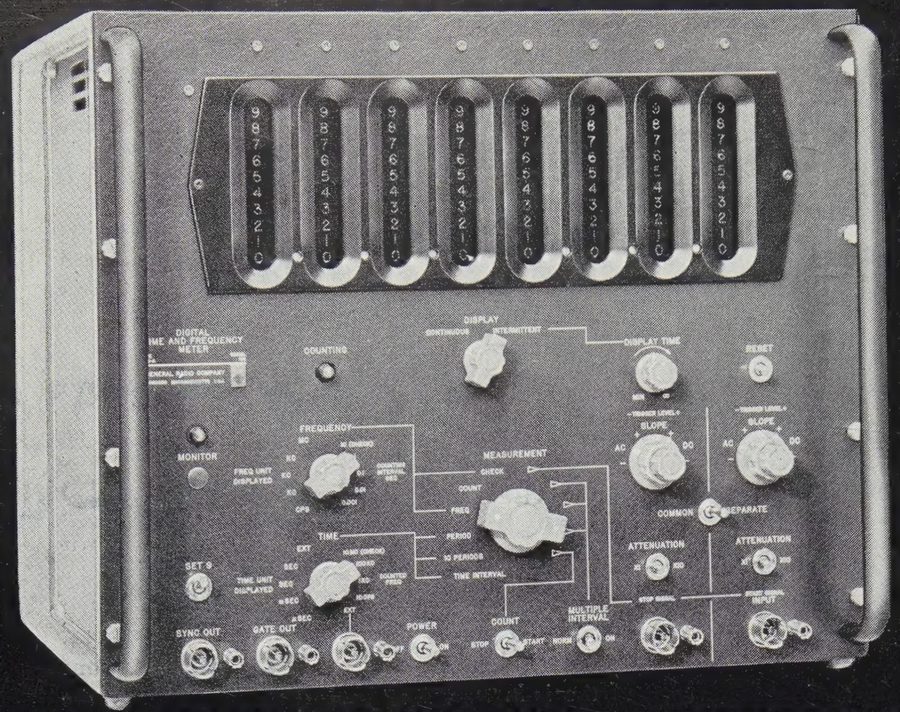

the Counter with a Memory

Continuous Readout to 10 Megacycles

The "memory" in this Counter constitutes an important new operating aid. Four of the instrument's eight decades are used for storage and continuous display, while the remaining four decades count continuously. At the end of each counting interval, the total accumulated by the counting decades is transferred automatically and quickly (only 100 μsec) to the storage and display decades. Continuous counting offers many advantages – information is sampled more often; frequency adjustments become easy; analog recording is greatly simplified; and operator eye fatigue induced by the dancing lights of intermittent displays is eliminated.

The Type 1130-A Digital Time and Frequency Meter is not just another counter. It embodies a number of new engineering contributions that are of fundamental importance.

This instrument is designed like a digital computer — to achieve a uniform level of high reliability throughout. "Down time", the bugaboo that robs the user of his full investment, is at a minimum.

Unsurpassed reliability is achieved by:

1. New decade codes and high-speed counting circuits, *unlike those in other counters*, that make this instrument inherently reliable.
2. Circuits designed to operate properly under the worst combination of *cumulative* tolerances imposed by tubes, component values, and voltage levels. Counter *performs properly even with tubes approaching the half-dead state.*
3. Use of proven "hard-bottoming" multivibrator dividers that make for exceptional stability — eliminate need for periodic adjustments of time-base circuits.
4. Elimination of critical voltages. Neither plate nor filament supplies are, or need be, regulated.

RANGES:
Frequency: dc to 10 Mc
Period: 10 μsec to 10^7 sec
Time Interval: 1 μsec to 10^7 sec
Also measures 10 periods, frequency ratios, phase shifts, pulse characteristics, and counts random events.

SENSITIVITY:
0.25v rms

DISPLAY
4 digits continuous; 8 digits for sequential counting and display, with display-time variable from 0.1 to 10 sec.

ACCURACY:
±1 count ± time-base oscillator stability

AVAILABLE WITH SEVERAL PLUG-IN TIME-BASE OSCILLATORS

Buy the Time-Base Stability You Need

	Complete Instrument Type - Price	Short-Term Stability Better Than	Long-Term Stability Better Than
Completely Self-Contained	1130-A4, $2,950.	1 part in 10^9 per min.	5 parts in 10^8 per week
	1130-A3, $2,670.	1 part in 10^8 per min.	2 parts in 10^7 per week
For Use from External Standards	1130-A2, $2,750.	——— Same as 1130-A3 ——— Also operates from external 100-kc, 1-Mc, and 5-Mc inputs.	
	1130-A1, $2,585.	Requires 5-Mc driving signal; G-R 1113-A 5-Mc Standard Frequency Oscillator provides stability of 1 part in 10^{10} per min, 2 parts in 10^9 per week.	

For Digital Recording

1132-A Data Printer . . . $1450.
Records 8 digits from counter plus 4 digits from clock or other source, at speeds to 3 prints per sec . . . no modification of counter is required

For Graphic Recording

1134-A Digital-to-Analog Converter . . . $595
Makes possible low-cost, ALL-ELECTRONIC graphic strip-chart recording (no data printer needed) . . . high accuracy of 0.1%

For Measurements to 500 Mc

Frequency conversion units are under development

Write for Complete Information

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